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VELOCITY-MODULATION VALVES FOR 100 TO 1000 WATTS CONTINUOUS OUTPUT

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In the generation of decimetric and centrimetric waves a great advance has been made by replacing the density modulation of the electron stream as used in transmitting valves by velocity modulation which leads at a further stage in the valve to density modulation. The interaction between the electric field and the electrons, which in valves of conventional construction has to take place over distances of tenths of a millimetre, can then take place over distances of some centimetres. This construction makes it possible to generate very much greater powers. In this article it will be explained how, with velocity-modulation valves for 3 to 15 cm wavelength, a continuous output of 100 to 1000 watts can be obtained.

Nowadays there are various special types of valves for generating and amplifying centrimetric waves. In addition to the triodes specially constructed for centrimetric waves — called disk-seal triodes, in which direct density-modulation of the electrons is applied — there may be mentioned the velocity-modulation valve ¹), the reflex klystron and the multireflex valve ²), and the travelling-wave tube ³). In these valves the electrons are velocity-modulated and this subsequently leads to density modulation. This principle is applied also in the magnetron, though in a somewhat more complicated manner ⁴). Whilst all these valves can be used as oscillators, the disk-seal triode, the velocity-modulation valve and the travelling-wave tube are also useful as amplifiers.

Which type of valve is to be employed depends entirely upon the requirements it has to meet for a particular purpose. As far as oscillators are concerned, to which this article will mainly be restricted, for small outputs (up to, say, 1 watt) reflex klystrons are used on account of their simple construction and operation, whilst also disk-seal triodes and special types of velocity-modulation valves are sometimes employed. For very high outputs (10 to 1000 kW), which are generated only in short pulses, exclusively magnetrons have been used hitherto. It is true that larger continuous outputs can be obtained with magnetrons, but these valves are difficult to tune and, moreover, the heating of the cathode by the returning electrons is apt to lead to instability. For this purpose, however, other suitable valves are available, such as the multireflex valve and the velocity-modulation valve.

The principles of the velocity-modulation valve have already been dealt with at some length in this journal ¹), and it was thereby pointed out that the theory expounded only roughly explained the operation of the valve. In this article, following

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¹) For a treatise on the principles of the velocity-modulation valve see F. M. Penning, Velocity-modulation valves, Philips Techn. Rev. **8**, 214-224, 1946.

²) F. Coeterier, The multireflection tube, a new oscillator for very short waves, Philips Techn. Rev. **8**, 257-266, 1946.

³) R. Kompfner, Wireless World **52**, 369-372, 1946. For the working of this valve as oscillator see also B. B. van Iperen, The helix as resonator for generating ultra-high frequencies, Philips Techn. Rev. **11**, 221-231, 1950.

⁴) A review of the working of various modern valves for centimetric waves is to be found in an article by F. M. Penning, Transmitting valves for decimetre and centimetre waves, Communication News **10**, 109-118, 1950.

a brief recapitulation of the principle of the valve, further consideration will be given to the practical realization of the principle and to some theoretical aspects of importance in the construction of the valves.

Elementary theory

The principle of the velocity-modulation valve can best be explained with reference to the diagram in *fig. 1*, where a stream of electrons is assumed to pass through two pairs of grids *a-b* and *c-d* in succession. Let the velocity with which the electrons reach the grid *a* be denoted by v_0 and the voltage required to give them this velocity by V_0. Now, when an alternating voltage $V_1 \sin \omega t$ is applied between the two grids *a* and *b* (the buncher or input gap), each electron is accelerated or retarded according to the instant at which it passes through this pair of grids. In the drift space, between the grids *b* and *c*, which are placed a fair distance apart, electrons which started later but have been accelerated between *a* and *b* have an opportunity to overtake other electrons which started earlier but which have suffered retardation. The result of this is that the density of the stream of electrons which at *a-b* was constant in time shows farther on periodical variations with the fundamental frequency ω. As the stream passes through the catcher or output gap between the pair of grids *c-d* its A.C. component induces an equally large alternating current in the circuit connected to the grids *c* and *d* (the output circuit). When this circuit is tuned close to the frequency ω, so that for that frequency it has a high impedance, then the stream of electrons imparts energy to that circuit. Thus the valve acts as an amplifier. For the voltage between the grids *a* and *b* to be induced with as little loss of energy as possible it is necessary to connect also these grids to a tuned circuit (input circuit). When this circuit is electromagnetically coupled with the output circuit an oscillator is obtained.

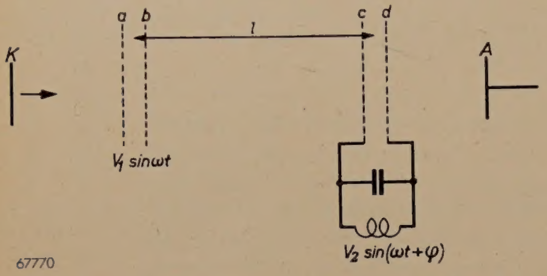

Fig. 1. Schematic representation of a velocity-modulation valve. The electrons emerging from the cathode *K* arrive with a velocity v_0 at the buncher or input gap bounded by the grids *a* and *b*, between which an alternating voltage is applied. Here the electron beam is modulated in velocity, so that in the drift space between the grids *b* and *c* the beam undergoes a density modulation. When passing through the catcher or output gap between the grids *c* and *d* the beam gives off energy to the output circuit coupled to those grids. Finally the electrons are collected at the anode *A*.

Since we shall be dealing here particularly with oscillator valves, it is especially important to consider the efficiency that can be reached with a construction such as that referred to above. First a distinction has to be made between the electronic efficiency, η_e, — that part of the D.C. power fed to the valve which is converted into H.F. power in the electron stream — and the overall efficiency η_t, defined similarly, but with the losses in the input and output circuits deducted from the H.F. power generated in the stream.

Assuming in the first instance that the circuit losses are negligible, we have only to deal with the electron efficiency, and for a construction such as that of fig. 1 this can be calculated in a fairly simple manner from the formula:

$$\eta_e = -\frac{V_2}{V_0} \sin(2\pi\xi + \varphi) \cdot J_1(\pi a\xi), \quad . \quad . \quad (1)$$

where V_2 is the amplitude of the alternating voltage between the catcher grids, φ is the phase difference between V_2 and the alternating voltage V_1 between the buncher grids, $a = V_1/V_0$ is the modulation depth, ξ the number of cycles of the H.F. alternating voltage required for an electron with velocity v_0 to traverse the distance from buncher to catcher, and J_1 is the Bessel function of the first kind and the first order.

Formula (1) is derived as follows. After having passed the buncher grids at a moment t_1, an electron has an energy

$$eV = eV_0 + eV_1 \sin \omega t_1 = eV_0 (1 + a \sin \omega t_1), \quad . \quad (2)$$

in which e is the charge of the electron. Taking v_0 and v_1 as representing the velocity of the electron before and after passing the grids, then $eV_0 = \frac{1}{2}mv_0^2$ and $eV = \frac{1}{2}mv_1^2$. Substitution in (2) gives:

$$v_1 = v_0 (1 + a \sin \omega t_1)^{\frac{1}{2}}.$$

Denoting the time of arrival of the electron at the second pair of grids by t_2 and the distance between the two pairs of grids by l, then

$$t_2 = t_1 + \frac{l}{v_0} (1 + a \sin \omega t_1)^{-\frac{1}{2}}.$$

Assuming that $a \ll 1$, thus that the depth of modulation is small, we may write:

$$t_2 = t_1 + \frac{l}{v_0} (1 - \tfrac{1}{2}\, a \sin \omega t_1). \quad . \quad . \quad . \quad . \quad . \quad (3)$$

If the alternating voltage across the catcher grids is $V_2 \sin(\omega t + \varphi)$ then, on passing these grids, the electron loses energy to an amount

$$E = -eV_2 \sin(\omega t_2 + \varphi),$$

which energy is imparted to the output circuit in the form of oscillation energy. Substituting (3) in this expression, we find for the energy given off by the electron:

$$E = -eV_2 \sin\left(\omega t_1 + \varphi + \frac{\omega l}{v_0} - \frac{\omega l a}{2v_0} \sin \omega t_1\right) \quad . \quad . \quad (4)$$

To arrive at the total amount of energy imparted to the output circuit per second, the values of E found for all electrons passing through in a period of one second have to be added. Thus the power yielded by the stream of electrons is:

$$P_{\mathrm{tot}} = -i_0 V_2 \cdot \frac{1}{2\pi} \int_0^{2\pi} \sin\left(\omega t_1 + \varphi + \frac{\omega l}{v_0} - \frac{\omega l a}{2v_0} \sin \omega t_1\right) \mathrm{d}(\omega t_1)$$

$$= -i_0 V_2 \sin\left(\varphi + \frac{\omega l}{v_0}\right) \mathrm{J}_1\left(\frac{\omega l a}{2v_0}\right),$$

where i_0 is the current density in the stream.

Now $\omega l/v_0 = 2\pi\xi$, and since $i_0 V_0 = P_0$ is the input power, the electronic efficiency is:

$$\eta_e = \frac{P_{\mathrm{tot}}}{P_0} = -\frac{V_2}{V_0} \sin(2\pi\xi + \varphi)\, \mathrm{J}_1(\pi a \xi).$$

The factors occurring in eq. (1) are now to be chosen such that η_e is the maximum obtainable. This means that the following conditions have to be satisfied:

(1) $\mathrm{J}_1(\eta a \xi)$ must be a maximum; such is the case for $\eta a \xi = 1.84$, whereby $\mathrm{J}_1(\eta a \xi) = 0.58$.

(2) V_2/V_0 has to be a maximum; this means that this quotient has to be made equal to unity, since for $V_2 > V_0$ some electrons when reaching the catcher grids would be turned back, in which case the foregoing theory no longer holds; a closer investigation shows that the efficiency is then greatly reduced.

(3) $\sin(2\pi\xi + \varphi) = -1$, thus, when n is a whole number, $2\pi\xi + \varphi = (n - \frac{1}{4})2\pi$, from which it follows that $\xi = n - \frac{1}{4} - \varphi/2\pi$.

Provided these three conditions are satisfied, the efficiency is 58%.

These conditions call for some further explanation. (3) can always easily be satisfied because ξ can be given any desired value by chosing a suitable initial velocity v_0 for the electrons. From condition (1) it follows that $a = V_1/V_0 = 1.84/\pi\xi$, from which the value of the modulation voltage is determined. Further, the output $P_{\mathrm{tot}} = V_2^2/2Z$, where Z represents the impedance between the catcher grids. If $P_0 = i_0 V_0$ (i_0 being the current in the beam of electrons) is the input power, then in the ideal case — when according to (2) $V_2 = V_0$ — the output is $P_{\mathrm{tot}} = 0.58\,P_0$, so that Z has to be given the value $V_0^2/1.16\,P_0 = V_0/1.16\,i_0$. For the case of an amplifier, therefore, with the proper choice of V_1, V_0 and Z, the maximum H.F. power can be derived from the beam of electrons.

In an oscillator valve, on the other hand, the value of V_1 cannot be directly adjusted. For such a valve the optimum adjustment is reached by a suitable choice of the feedback factor, i.e. the ratio $K = V_1/V_2$, which can indeed be directly adjusted, as will presently appear. For adjustment to the maximum electronic efficiency again in the ideal case $V_2 = V_0$, and thus we find: $K = V_1/V_2 = V_1/V_0 = a = 1.84/\pi\xi$.

In practice, however, it is not so much a mattar of the maximum electronic efficiency but rather the maximum overall efficiency. In the first place the losses in the input and output circuits cause the overall efficiency η_t to be smaller than the electronic efficiency η_e. If the buncher and the catcher have the same impedance Z_k then the circuit losses amount to $(V_1^2 + V_2^2)/2Z_k$. Therefore, to obtain a high overall efficiency it is of importance to make Z_k as large as possible.

Moreover, the maximum of η_t need not occur with the same value of the feedback factor K as does the maximum of η_e.

This is readily understood when considering the condition under which the valve just begins to oscillate. In that case the output power from the electron stream is just equal to the circuit losses, so that

$$i_0 V_2 \mathrm{J}_1(\pi a \xi) = (V_1^2 + V_2^2)/2Z_k,$$

and since with a small amplitude $\mathrm{J}_1(\pi a \xi) \approx \frac{1}{2}\pi a \xi$, this becomes:

$$\tfrac{1}{2} i_0 V_2 \left(\pi\xi \frac{V_1}{V_0}\right) = V_2^2 \frac{1 + K^2}{2Z_k},$$

or

$$i_0 = \frac{1 + K^2}{K} \frac{V_0}{\pi\xi Z_k}.$$

This expression is a minimum when $K = 1$, in which case

$$i_0 = \frac{2V_0}{\pi\xi Z_k}. \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

For $K = 1$, therefore, the current at which the valve begins to oscillate is lowest. Just above this minimum current determined by (5) the valve therefore oscillates only when K is made equal to 1, at which value in this case also the (here very small) overall efficiency is a maximum. For the value of K at which the electronic efficiency is maximum we found, on the other hand, $K = 1.84/\pi\xi$, which usually is much less than unity. In general the value of K for a maximum total efficiency will lie between 1 and $1.84/\pi\xi$.

Adjustment of the feedback

Adjustment of the feedback factor — the quantity $K = V_1/V_2$ — of two coupled circuits is not a special problem typical of the velocity-modulation valve but also arises with a triode oscillator working with two tuned *L-C* circuits. The behaviour of such systems has already been thoroughly investigated [5]), but since the adjustment of the feedback factor is less generally known this problem will be briefly explained here.

Let us assume that the input and output circuits are inductively coupled (*fig. 2*) by a mutual inductance M. As is well known in radio engineering, the behaviour of such a system depends upon the coupling coefficient $f = M/\sqrt{L_1L_2}$ (not to be confused with the feedback factor K), the quality factors Q_1 and Q_2 and the two natural frequencies ω_1 and ω_2 of the separate circuits. For the case in which we are interested we may for the time being assume that $1/Q_1 = 1/Q_2 = \delta$, while it is further assumed that $\delta \ll f \ll 1$.

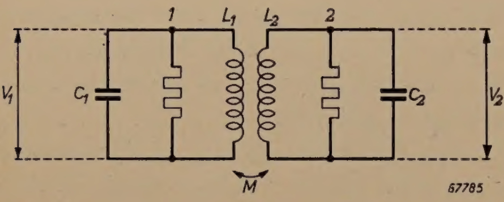

Fig. 2. Two inductively coupled circuits representing schematically the input circuit (*1*) and the output circuit (*2*) of a velocity-modulation valve used as oscillator. With a suitably chosen difference between the natural frequencies of the two circuits it is possible to adjust the feedback factor V_1/V_2 to any desired value, provided the coupling coefficient $f = M/\sqrt{L_1L_2}$ between the circuits is large enough.

Now three cases are to be considered:

(1) The naturel frequencies of the two circuits are equal, thus $\omega_1 = \omega_2$.

(2) The relative difference between the natural frequencies is very great compared with the coupling coefficient, thus $|\omega_1 - \omega_2|/\omega_1 \gg f$.

(3) The relative difference between the natural frequencies is of the same order as the coupling coefficient.

In the case where $\omega_1 = \omega_2 = \omega_0$ the system has two natural frequencies: $\omega_a = \omega_0\,(1 + f)$ and $\omega_b = \omega_0(1 - f)$. At the frequency ω_a the voltages V_1 and V_2 are in phase, thus $\varphi = 0$, whereas at the frequency ω_b they are in anti-phase ($\varphi = 180°$). Further, from considerations of symmetry it follows that in both cases $V_1 = V_2$, so that the feedback factor $K = 1$. This can be represented schematically as in *fig. 3*.

In the second case, where $|\omega_1 - \omega_2|/\omega_1 \gg f$, there are also two natural frequencies but there they are almost equal to the natural frequencies of each of the circuits separately. At the natural frequency $\omega_a \approx \omega_1$ the value of V_1 becomes large, while the oscillation of the second circuit is only weak, so that V_2 remains small ($K \gg 1$). At the other natural frequency, $\omega_b \approx \omega_2$, the situation is just the reverse ($K \ll 1$). The phase relations remain the same as in the first case. Schematically this is as represented in *fig. 4*.

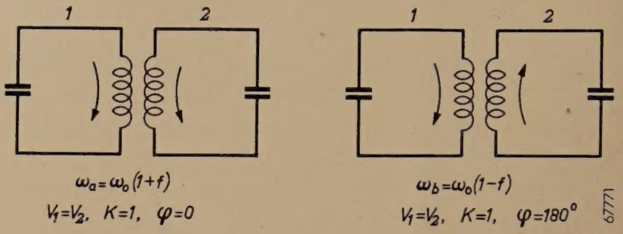
1
2
$\omega_a = \omega_0(1+f)$
$V_1 = V_2$, $K=1$, $\varphi=0$
1
2
$\omega_b = \omega_0(1-f)$
$V_1 = V_2$, $K=1$, $\varphi=180°$
67771

Fig. 3. Schematic representation of the mode of oscillation of the system according to fig. 2, for the case where the two circuits are tuned to the same frequency. At the higher of the two natural frequencies the two circuits are in phase, and at the lower one they are in anti-phase. In both cases the feedback factor K is equal to unity.

In a velocity-modulation valve usually K is required to be less than unity, but not very small. This can be achieved by choosing for $|\omega_1 - \omega_2|$ a value between those of the extreme cases. This brings us to the third case mentioned above, where $|\omega_1 - \omega_2|/\omega_1$ is of the same order as f. The manner in which K depends upon the detuning $(\omega_1 - \omega_2)/\omega_1$ between the two circuits is shown in *fig. 5*, from which it appears that any desired value of K can be reached by a suitable choice of $\omega_1 - \omega_2$ (provided, however, that the above-mentioned condition that the coupling coefficient f is large compared with $1/Q = \delta$, is satisfied).

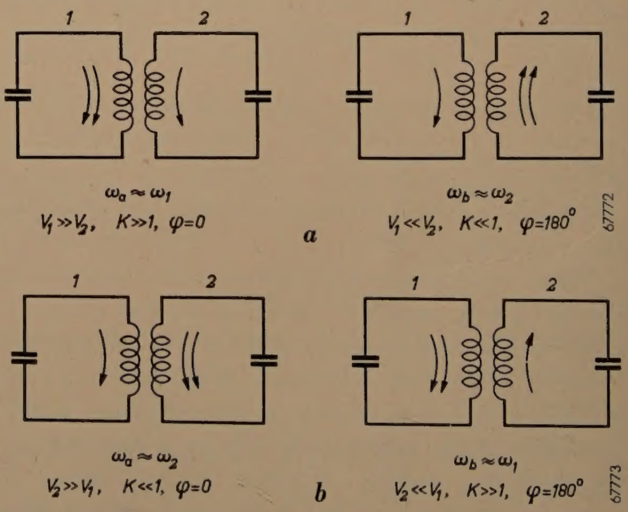
1
2
$\omega_a \approx \omega_1$
$V_1 \gg V_2$, $K \gg 1$, $\varphi=0$
a
1
2
$\omega_b \approx \omega_2$
$V_1 \ll V_2$, $K \ll 1$, $\varphi=180°$
67772
1
2
$\omega_a \approx \omega_2$
$V_2 \gg V_1$, $K \ll 1$, $\varphi=0$
b
1
2
$\omega_b \approx \omega_1$
$V_2 \ll V_1$, $K \gg 1$, $\varphi=180°$
67773

Fig. 4. Schematic representation of the mode of oscillation of the system according to fig. 2 when the natural frequencies of the circuits differ greatly. Here again there are two natural frequencies, but at one of these $K \gg 1$ (very high feedback factor) whilst at the other $K \ll 1$ (very low feedback factor). In the case (*a*) the natural frequency of the first circuit is much greater than that of the second one, in the case (*b*) it is much smaller.

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[5]) See, e.g., B. D. H. Tellegen, Philips Res. Rep. 2, 1-19, 1947.

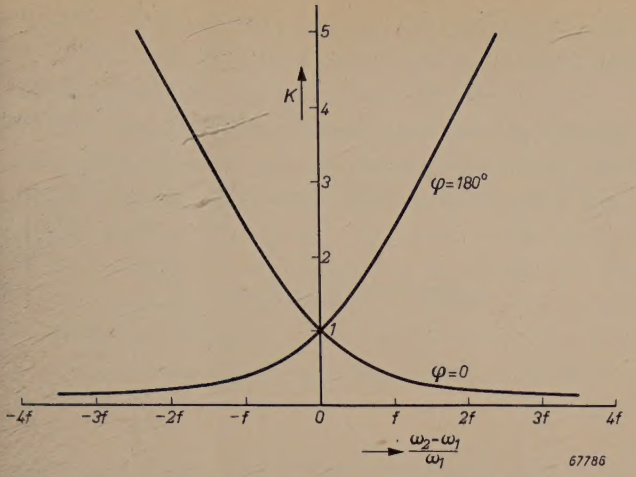

Fig. 5. Relation between the feedback factor K and the detuning of the circuits, for the system of fig. 2. For each value of the relative detuning $(\omega_2 - \omega_1)/\omega_1$ there are two natural frequencies, corresponding to the two curves in the diagram. When with one mode of oscillation, e.g. that where the circuits are in phase ($\varphi = 0$), K is greater than 1, then with the other mode of oscillation ($\varphi = 180°$) K is less than 1, and vice versa. f is the coupling coefficient between the circuits.

General principles of the construction

These considerations are useful for a general insight into the working of a velocity-modulation valve. When, however, the principle is applied for generating energy at a very short wavelength it appears that the conception from which we started is quite inadequate, because account has to be taken of factors which have not yet been included. This can best be explained by first examining more closely the practical construction of a velocity-modulation valve.

If, as was supposed, grids were used in the path of the electron beam then these could have only a small surface, because in the oscillator circuit they form a capacitance which adversely affects the resonance impedance. Owing to the bombardment by the electrons these grids reach a very high temperature, so that the input power has to remain extremely small. For high outputs, therefore, it is preferable to choose a solution whereby each pair of grids is replaced by two hollow cylinders placed in line with each other and with a gap between them. These are respectively the input gap and the output gap. For the oscillator circuits use is made of cavity resonators of such a shape and of such a material that the highest possible impedance is reached.

The valve has to answer a number of other requirements too. In the first place it must be possible to vary the natural frequency of the cavity resonators, since the valve should preferably be tunable, and moreover, as already seen, with a view to adjusting the value of the feedback factor K the natural frequencies of the two cavity resonators should be adjustable with respect to each other. Further, in order to obtain the highest possible output, the tube should be capable of large dissipations.

These latter requirements lead to the valve being so designed that the wall of the cavity resonators serves at the same time as the vacuum envelope. *Fig. 6* gives a cross-sectional (schematic) view of

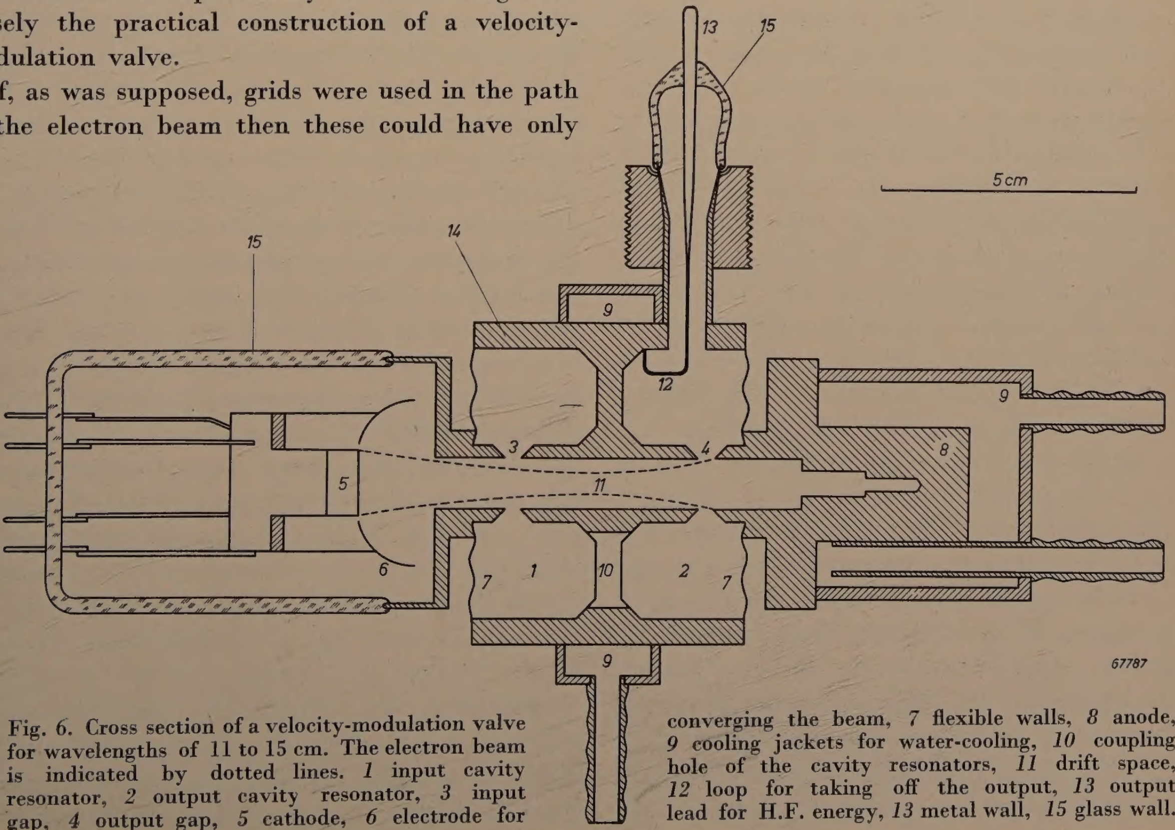

Fig. 6. Cross section of a velocity-modulation valve for wavelengths of 11 to 15 cm. The electron beam is indicated by dotted lines. *1* input cavity resonator, *2* output cavity resonator, *3* input gap, *4* output gap, *5* cathode, *6* electrode for converging the beam, *7* flexible walls, *8* anode, *9* cooling jackets for water-cooling, *10* coupling hole of the cavity resonators, *11* drift space, *12* loop for taking off the output, *13* output lead for H.F. energy, *13* metal wall, *15* glass wall.

a valve constructed in this way; for a detailed explanation see the subscript. The shape of the electron beam is indicated by dotted lines. The natural frequency of the cavity resonators can be varied by changing the width of the gaps, this being possible owing to one of the side walls of each of the cavity resonators being made thin and thus flexible. Since the gap edges form the major part of the capacitance of the cavity resonator, by only slightly changing the gap width a considerable change in wavelength can be obtained.

Effects not taken into account in the elementary theory

As we have seen, under certain assumed conditions the simple theory predicts an efficiency of 58% when disregarding the losses in the cavity resonators, which in practice may well be considerable. In the construction of a valve as described above it is not easy to realize to a good approximation the conditions assumed in the elementary theory. To make this clear we have to consider more closely the manner in which the electrons traverse the system and the interaction between the electron beam and the electromagnetic field.

Influence of mutual repulsion of the electrons

So far it has been assumed that the electrons travel in straight paths, and this is indeed possible of attainment to a good approximation when a very strong magnetic field is placed in the direction in which the electrons are travelling. Such a solution, however, is rather impracticable and preference is therefore given to a valve without magnetic field. Under the influence of their mutual repulsion the electron paths which are originally parallel in a beam will begin to diverge, the more so the stronger the current at a particular electron velocity. Consequently, with a cylindrical tube of a given diameter and length, and with a certain initial velocity of the electrons, a limit is set to the current that can be sent through the tube without causing electrons to be lost at the wall of the tube.

It is better, therefore, to start with a convergent beam of electrons. A convergent beam reaches a minimum diameter at a certain point and then begins to diverge; from the theory it follows that the shape of the beam is symmetrical with respect to this minimum. It is easily seen that a beam with the greatest possible current can be sent through a given tube when the minimum diameter of the beam is half-way along the length of the tube (*fig. 7*). From the theory it follows further that the maximum current which can pass through the tube in this way is proportional to the square of the ratio of the tube diameter to the tube length.

The theory outlined above shows that it is desirable to have the highest possible beam current for a given voltage. In order to maintain the necessary voltages across the input and output circuits it is necessary to supply the power equivalent to the circuit losses. Increasing the current then increases the H.F. power in the beam and this increase is entirely available as output. Thus the efficiency of the tube increases with the beam current.

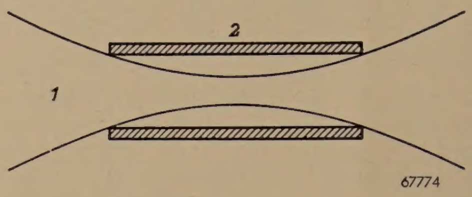

Fig. 7. Owing to mutual repulsion of the electrons, a beam which initially is convergent passes through a minimum diameter and then diverges. The shape of the beam is symmetrical with respect to the minimum. When the beam is sent through a round tube of a certain length and diameter it can be given the greatest current strength without electrons colliding with the wall by ensuring that the minimum diameter of the beam is reached half-way along the tube. *1* = electron beam, *2* = round tube.

Influence of the gap field on the paths of the electrons

The divergence of the beam is not the only factor restricting the current strength. Owing to the deflecting action of the H.F. field in the input gap we have to keep well below the limit just mentioned. To explain this let us assume for the sake of simplicity that before reaching the bunching field the beam is parallel. Now let us follow an electron travelling along a path at a certain distance from the axis of the tube (*fig. 8*) and assume that it passes the centre of the field at the moment when the alternating voltage between the gap edges has the maximum value. The radial force to which this electron is subjected during the first half of its passage through the gap will then be equal and opposite to that acting upon the electron during the second half of its passage. To a first approximation, therefore, this electron does not undergo any change in direction (path *1*, fig. 8). The same applies when the H.F. field has its maximum negative value.

Now suppose, however, that the electron passes through the centre of the gap just when the alternating voltage is passing through zero from "accelerating" to "retarding". During the first half of the time spent in the gap the electron is deflected towards the axis, and because of the reversal of field the deflection will be in the same direction during the second half of the transit. Thus this electron undergoes a change of direction towards the axis (path *2*).

The reverse applies to an electron passing through the field when the voltage is passing through zero from "retarding" to "accelerating". Such an electron is deflected outward, the more so the nearer its path lies to the outer edge of the beam, i.e. the closer the electron passes along the gap edges (path *3*). This deflection may easily reach such proportions that the electron strikes against the wall of the drift space before it has passed the output gap.

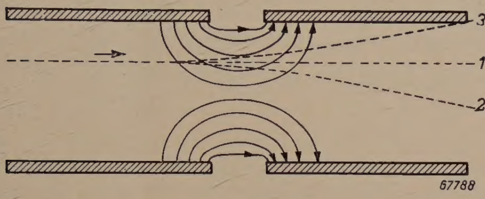

Fig. 8. An electron passing through the H.F. field in the gap of the input cavity undergoes a deflection which varies according to the time that the electron passes that gap.
1 path of an electron passing at the instant that the field has its maximum accelerating or retarding force.
2 path of an electron passing at the instant that the field is changing from the accelerating to the retarding phase.
3 path of an electron passing at the instant that the field is changing from the retarding to the accelerating phase.

Now this is most unfortunate, since these outwardly deflected electrons happen to be the most useful. This can be explained as follows. Electrons passing through just in front of the one we have been considering here have undergone a retardation (negative velocity modulation) and electrons immediately following it undergo an acceleration. On their way to the inductor the electrons of the group in question thus draw closer together and form a so-called bunch. The valve works best when this bunch of electrons enters the output gap at the moment that the H.F. voltage is in its maximum retarding phase, so that the group of electrons in question undergoes a strong retardation. If, owing to deflection in the input gap, part of this group of electrons strikes against the wall of the drift tube, then naturally this reduces the efficiency of the valve. For this reason care has to be taken to ensure that when passing the input gap the beam is fairly narrow. It will presently be shown how this undesired effect can be further counteracted by a special construction of the output gap.

Energy exchange between gap field and electrons

From the foregoing it would seem that the tubes through which the electron beam is sent should be given the largest possible diameter, that is to say, as large as is compatible with the construction of a cavity resonator with a suitable impedance at a certain frequency.

It appears, however, that on other grounds we are very limited in the choice of the tube diameter. To explain this it is necessary to reexamine the elementary theory. There we assumed a pair of grids with an alternating voltage $V_1 \sin \omega t$, and when calculating the electronic efficiency it was assumed that an electron, after passing through the pair of grids with an energy eV_0 at the instant t_1, had an energy $e(V_0 + V_1 \sin \omega t_1)$. This is only exact if the transit time of the electron between the two grids is very short compared with the cycle of the H.F. voltage. If that is not the case then the energy transferred to the electron is always less than $eV_1 \sin \omega t$. It may be proved that for $V_1 \ll V_0$, the energy variation can always be represented as $\beta e V_1 \sin \omega t_1$, where β is a constant smaller than unity. This applies not only for the case where we have a field between two grids but also for the case where the electron passes through a field of a different form, as for instance the H.F. field in the input gap of a valve constructed as in fig. 6. For this latter case, however, β is dependent on the distance between the path of the electrons and the axis of the tube.

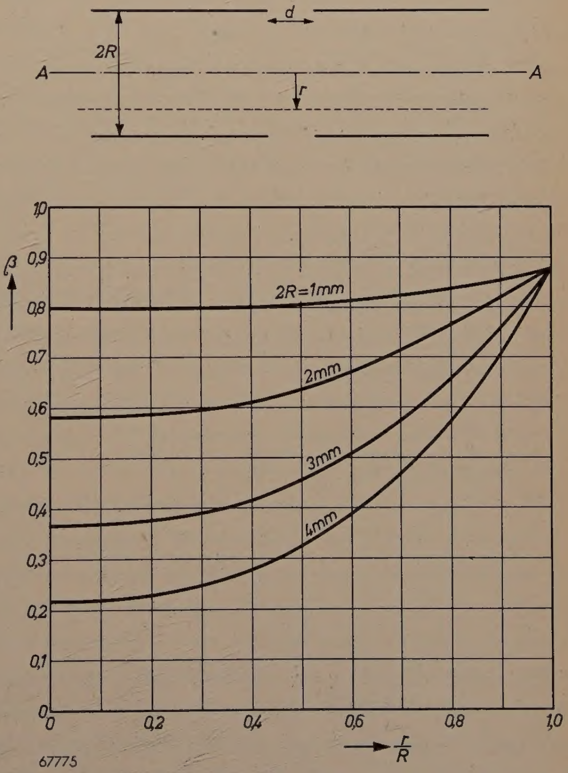

Fig. 9. Example of the variation of the gap factor β as a function of the distance r from the electron path to the axis *AA* of the tube. The diagram applies for the case of an accelerating voltage V_0 = 6000 V, a wavelength λ = 3 cm, a gap width d = 0.5 mm and for different tube diameters $2R$.

An example of the variation of β is to be found in *fig. 9*, from which it is seen that for a not too small value of β one is limited to small tube diameters. The importance of a high value of β can easily be understood: if the energy of the electron is to be changed by an amount eV_1 by means of the H.F. field then this requires an alternating voltage V_1/β, thus a power V_1^2/β^2Z_k. Something similar applies also for the output gap. It may therefore be said that the effect in question apparently reduces the impedance of the circuits by a factor β^2.

The fact that β depends upon the distance of the electron path from the axis of the tube has also an adverse effect upon the efficiency of the valve, in that the electrons do not all undergo the same velocity modulation, so that the modulation voltage cannot be made optimum for all electrons in the beam at the same time. The same holds in respect of the interaction with the output circuit, the electrons in the middle of the beam always giving off less energy than those on the outside.

Gap damping

Another phenomenon also arising as a result of the finite transit time of the electrons in the H.F. field is gap damping. By this is meant the following. When an electron beam passes through two grids placed close together and modulating the velocity of that beam, on an average the modulating field does not impart any energy to the beam, since just as many electrons are accelerated as are retarded, whilst the average acceleration equals the average retardation. It appears, however, that this no longer strictly holds if the transit time in the H.F. field is not very short compared with the time of oscillation. In most cases a certain amount of energy is then imparted to the beam by the buncher. This, therefore, constitutes a damping of the input circuit, i.e. an apparent reduction of the impedance of that circuit. Something similar is also the case with the output circuit.

Finally it should be pointed out that in the elementary theory it was assumed that the modulation depth $a \ll 1$, but in practice this is not correct either.

The result of all these effects is that the overall efficiency of velocity-modulation valves remains very much below the calculated value of 58%. We shall not here go further into the question as to how a more exact theory should be formulated. Suffice it to say that in practice an overall efficiency of 25% is to be regarded as very satisfactory.

The electron gun

From the discussion of the static electron paths we have seen that the beam entering the oscillator system is required to have a certain angle of convergence. A method by which such a beam is obtained has been suggested by Pierce 6) and will be briefly described as follows.

First imagine that we have a diode consisting of two concentric spheres with the outer one acting as a cathode. When a positive voltage is applied to the inner sphere then, for reasons of symmetry, the electrons coming from the cathode move towards the anode in straight lines. Let us now focus our attention upon a circular part of the cathoden surface, from which a convergent beam of electrons is moving towards the anode (see *fig. 10*). If we remove the other electrons, for instance by making the rest of the cathode surface non-emitting, then the rectilinear motion of the electrons is disturbed, since the potential distribution in the space between cathode and anode has been changed. Now it is possible to make the electrons in the cone still follow the original straight paths by making, in some way or other, the potential distribution along the conical surface again equal to the original distribution. This is done by setting up metal electrodes of a suitable shape close to the beam. The shape of these screens can be determined experimentally in an electrolytic tank, and they may also be

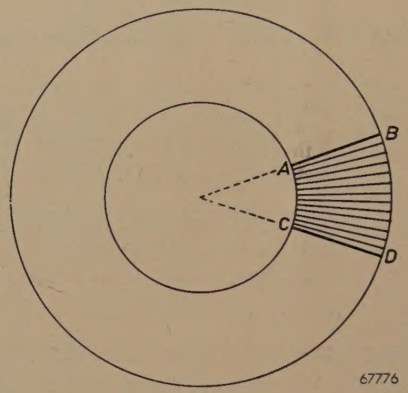

Fig. 10. In a diode consisting of two concentrical spheres, the outer one of which acts as a cathode, the electrons follow straight paths. If only a circular piece *BD* of the outer sphere is permitted to emit electrons then the electron paths within the cone *ABCD* remain rectilinear provided steps are taken to ensure that the potential distribution along the conical surface is the same as that obtaining when the whole of the outer sphere is emitting.

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6) J. R. Pierce, J. appl. Phys. 11, 548-554, 1940. The fact that the problem here is quite different from that in other valves in which narrow electron beams are used, such as cathode-ray tubes, is due to the much higher beam current required.

calculated 7)). In practice it is found to be sufficient if two electrodes are used, the first one at about cathode potential and the second at anode potential, the shape of the latter electrode not being very critical. As second electrode, therefore, the front of the oscillator system, where the beam enters, serves the purpose.

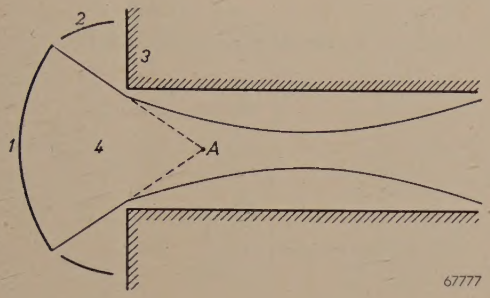

Fig. 11. Shape of the electron beam in a non-oscillating velocity-modulation valve (for the sake of clarity the transverse dimensions are exaggerated). A conical electron beam *4* (converging towards *A*) leaves the cathode *1* provided with a screen *2*. Owing to the divergent action of the hole in the oscillator system *3* the beam becomes less convergent, reaches its minimum diameter half-way along the system and then begins to diverge.

It is to be noted that the beam emerging from the cathode has to be more convergent than that ultimately required, since the opening through which the beam enters the oscillator system acts as a negative lens. Provided there are no oscillations, the shape of the beam is then as represented in *fig. 11*, where its width has been exaggerated for the sake of clarity. In the valve the screen round the cathode is fixed to a can in which the cathode is mounted in the right position with respect to that screen. In valves designed for wavelengths below 10 cm this unit is fixed, and insulated, to a ring fitting in a recess on the front of the oscillator system. In this way the electron gun — the combination of cathode and screen — can be exactly centered with respect to the oscillator system. A photograph of some guns for the various valves to be discussed is given in *fig. 12*.

The cathode

The fact that electron guns can be constructed in the simple way described above is due entirely to the use of the L-cathode, a new type of cathode already fully described in this journal 8)). At first

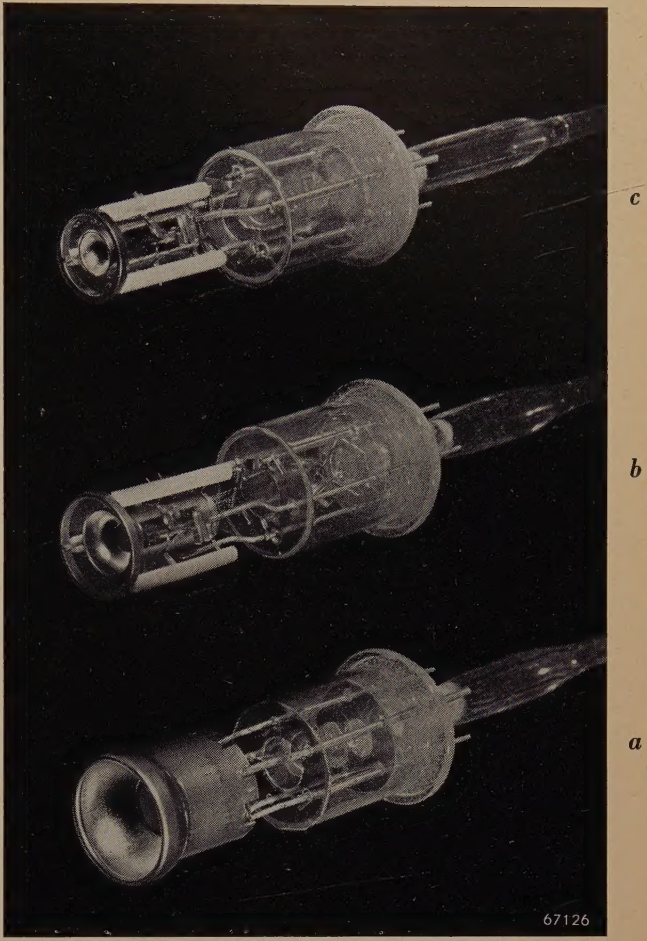

Fig. 12. Photograph of some electron guns as used in the valves described in this article. From *bottom* to *top*: *a*) gun of a valve for wavelengths of 11 to 15 cm and of a valve for 10 cm, 1200 watts; *b*) gun of a valve for 5.5 cm to 7 cm; *c*) gun of a valve for 2.9 cm to 3.6 cm. The glass cylinders between the gun and the bridge prevent the getter (barium) from being precipitated on undesired places.

the construction of the cathode was a difficult problem, because a valve of optimum dimensions involves comparatively high voltages (for one of the types to be described later, even as high as 15 kV), whilst moreover for smaller wavelengths the beam diameter, and thus the cathode diameter, has to be made smaller and smaller, resulting in great current densities (up to 2.4 A/cm^2 for the smallest type constructed).

It is therefore not possible to use the normal oxide-coated cathode in the gun construction described here. Owing to the high tension alone the inevitable bombardment with positive ions (formed from residual gases) would destroy the oxide coating of the cathode within a few hours. Apart from that, the current density required is too great for the cathode to last a reasonable length of time.

There are, it is true, various techniques for circumventing such difficulties. In larger types of

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7) According to a method indicated by E. G. Dorgelo (not published).

8) J. H. Lemmens, M. J. Jansen and R. Loosjes, A new thermionic cathode for heavy loads, Philips Techn. Rev. **11**, 341-350, 1950.

valves, for instance tungsten and tantalum have been used as cathode material, but owing to the high working temperature the heating of these metals gives rise to great difficulties which can only be overcome by rather complicated constructions. (It is to be borne in mind that the cathode has to be a circular, almost flat, disc.) A second objection is that these cathodes require a high heater power. Furthermore, this solution is not practicable in valves for very small wavelengths owing to the great current densities required.

The L-cathode, on the other hand, fully meets the requirements. Notwithstanding the great densities of current it has a satisfactory life, and it can also withstand bombardment with high-velocity ions. It needs heating to a temperature only slightly higher than that of a normal oxide-coated cathode, so that a comparatively low heater power suffices.

Feedback and tuning of the cavity resonators

What has been said in the foregoing about the feedback in the case of *L*-*C* circuits applies in the main also for two coupled cavity resonators. Here the coupling is brought about via a hole in the wall between the cavities, and since in velocity-modulation valves this hole is situated at a point where there is mainly a magnetic alternating field, this corresponds to an inductive coupling. The coupling coefficient f is fixed by the size of the coupling hole. In practice f is about $\frac{1}{2}$%, whilst for the unloaded cavity resonators $Q = 1/\delta$ is about 2000 to 4000, so that the condition $f \gg \delta$ is satisfied. It is true that the quality factor of the output cavity resonator is reduced by the loading, but this does not essentially change the behaviour of the system in regard to the variability of the feedback factor.

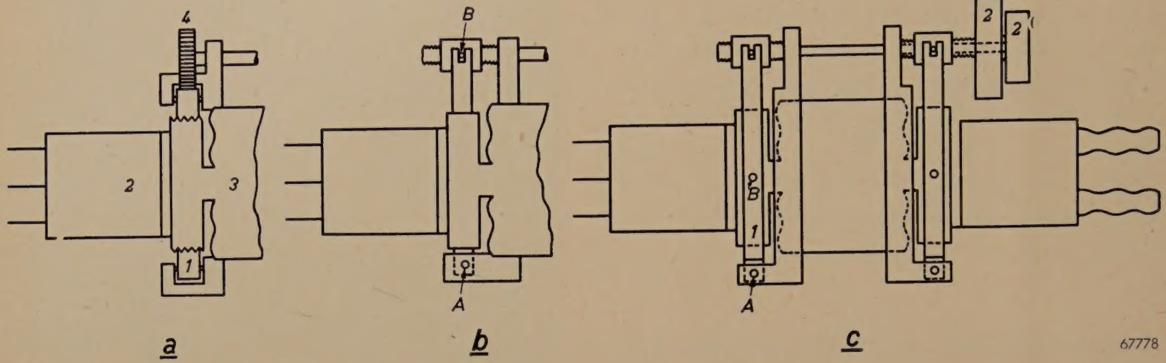

Fig. 13. *a*) Schematic representation of a tuning device operated with the aid of a threaded ring fitting round the movable part of the tube. Rotation of the ring imparts an axial movement to the cathode, respectively anode, part of the valve (here only the cathode end of the valve is shown). *1* ring toothed on the outside and rotated by a small gear wheel *4*. *2* cathode part, *3* middle part of the valve.
b) Tuning device working on the lever principle. The cathode part is pivoted to the middle part at *A* and moved by some means or other at *B*. Here the movable part describes a circular motion instead of an axial one.
c) New tuning mechanism consisting of a ring (*1*) pivoted to the middle part at *A* and likewise pivoted to the cathode part at *B* (two diametrically opposed points). The motion is now axial to a sufficient degree of accuracy. The rings are moved by turning two knobs *2* each serving a cavity resonator. The spindle for one resonator has left-hand thread and that for the other right-hand thread. See further fig. 14.

For adjusting the feedback factor and the wavelength of the oscillations it must be possible to vary the frequency of each of the cavity resonators. To that end provision has to be made for accurately adjusting the width of the gaps in the input and output cavities. It has already been shown in fig. 6 how the parts of the cavities at the cathode and anode ends of the valve are connected with the middle of the valve by means of a diaphragm. Thus the valve consists of three parts: the cathode, the anode and the middle parts, which have to be movable axially with respect to each other.

An obvious means of bringing this about is indicated in *fig. 13a* (only the cathode end has been drawn, the situation being the same at both ends). A threaded ring is fitted round the part to be moved, the position of this ring with respect to the middle part being fixed by means of a groove, so that when the ring is turned the cathode part is displaced with respect to the middle part. Such a device, however, is difficult to make without leaving some play, and there is much friction.

Another frequently applied device, with lever action, is represented in fig. 13*b*. This is very simple but it has the drawback that the slope of the moving part with respect to the middle part of the valve is changed, so that this system can only be used for very small variations of the gap width.

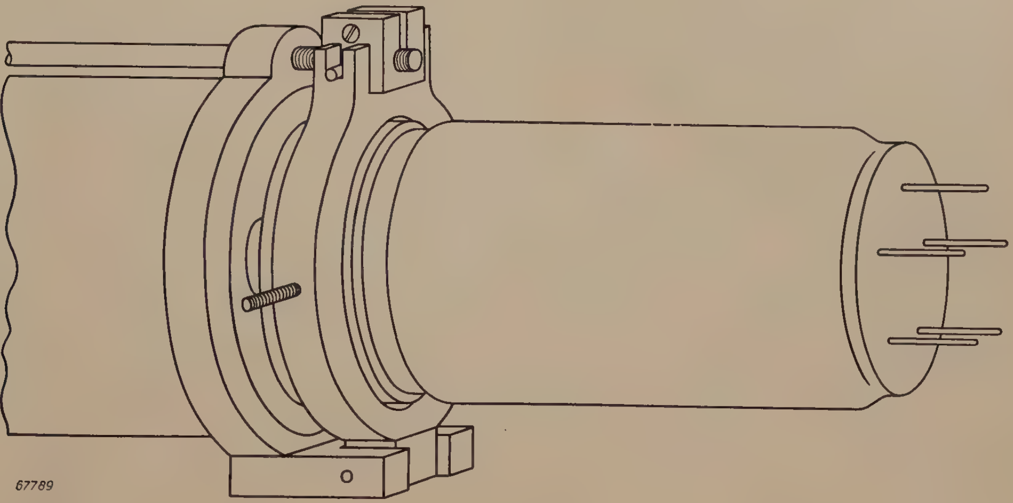

Fig. 14. Sketch showing one end of a valve with the mechanism according to fig. *1*3*c*.

The construction applied for the valves to be described combines the advantages of the two systems just mentioned (see fig. 13*c* and *fig. 14*). Here the lever is in the form of a ring placed round the movable cathode part and connected with it at two pivoting points. The ring is likewise pivoted at the bottom in a junction piece connecting it with the middle part of the valve. Thus a movement is obtained which is of sufficient axial accuracy for the desired range. The two cavity resonators are operated by means of two knobs mounted on concentric spindles, as shown in fig. 13*c*. Right-hand thread is used for the adjustment of one resonator and left-hand thread for the other, so that when both knobs are turned in the same direction the frequency variation brought about in the two cavities is also in the same sense, which is most convenient.

Improved construction of the output gap

When studying the paths followed by the electrons we saw that in the buncher space the electrons underwent a change of direction, such that the most useful electrons were deflected outward and might well strike against the wall of the drift tube and thus be lost.

In some of the types of valves to be mentioned later, this possibility has been greatly reduced in the following way. The electrons are allowed to undergo deflection in the H.F. field in the manner described, so that without further steps being taken they would impinge on the wall of the drift tube. In this wall, however, and also in that which bounds the catcher field of the other side, longitudinal

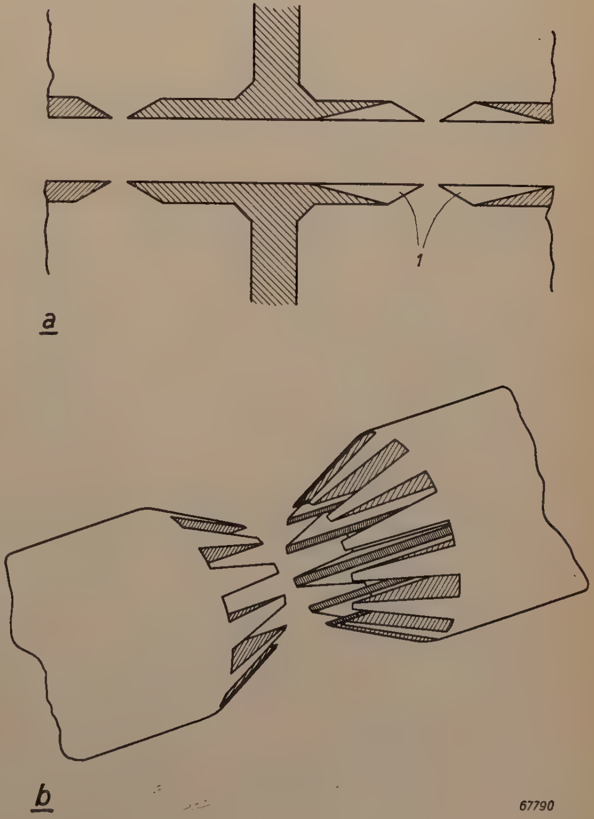

Fig. 15. *a*) Sketch of the output gap in a valve provided with longitudinal grooves *1* in the wall of the drift tube in order to increase the effective beam current. *b*) Schematic representation of the shape of the grooves in a valve as in (*a*).

grooves have been made through which the electrons can for the greater part proceed further on their way and pass through the catcher field. As *fig. 15* shows, the edges of the output gap are thus given the shape of a toothed ring. With this construction it appears to be possible to increase the beam current considerably and at the same time to improve the efficiency of the valves, so that with the same voltage a much larger H.F. output can be obtained.

Some experimental types of valves

Four experimental types of valves have been constructed in the Philips Laboratory.

A cross-sectional diagram of one of them has already been given in fig. 6. For the second type the essential dimensions of the valve were halved, thereby making it suitable for frequencies twice as high. These two types have both been made with the walls of the output gap grooved in the manner just described, so as to increase the effective beam current. The first type of valve is suitable for generating wavelengths of 11 to 15 cm, the greatest output — about 500 watts — being obtained at about 12 cm. With this adjustment the valve voltage is 9 kV and the beam current 220 mA, giving an overall efficiency of 25%. As the wavelength is increased the voltage has to be reduced in proportion to the square of the frequency. Also the beam current then drops, so that the output diminishes rather tapidly. In this range the valve works with an elecrron transit time ξ between the input and the output gaps of $1\frac{3}{4}$ cycles. At wavelengths below 12 cm the electrons can be given a transit time of $2\frac{1}{4}$ cycles. The output of this valve is via a coaxial terminal, the inner conductor of which passes out through a glass cap. A coaxial line can be connected to this system.

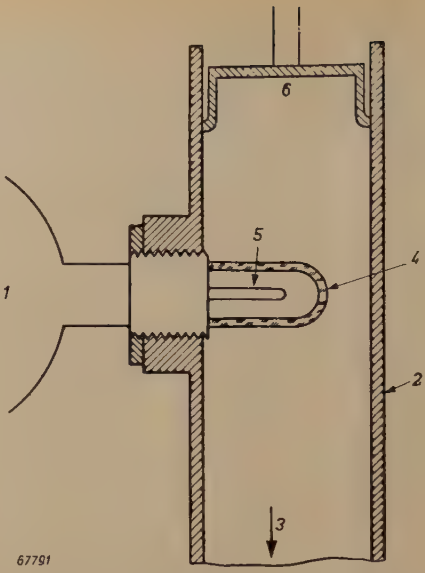

Fig. 16. Schematic representation of the device for extracting energy from a velocity-modulation valve by means of a coaxial output terminal protruding into a rectangular wave guide. *1* valve, *2* wave-guide, *3* direction in which the energy is conducted, *4* glass cap, *5* inner conductor of the coaxial output terminal, *6* plunger.

The second type is suitable for wavelengths of 5.5 to 7 cm, giving at 6.1 cm a maximum output of over 300 W (voltage 8.5 kV, beam current

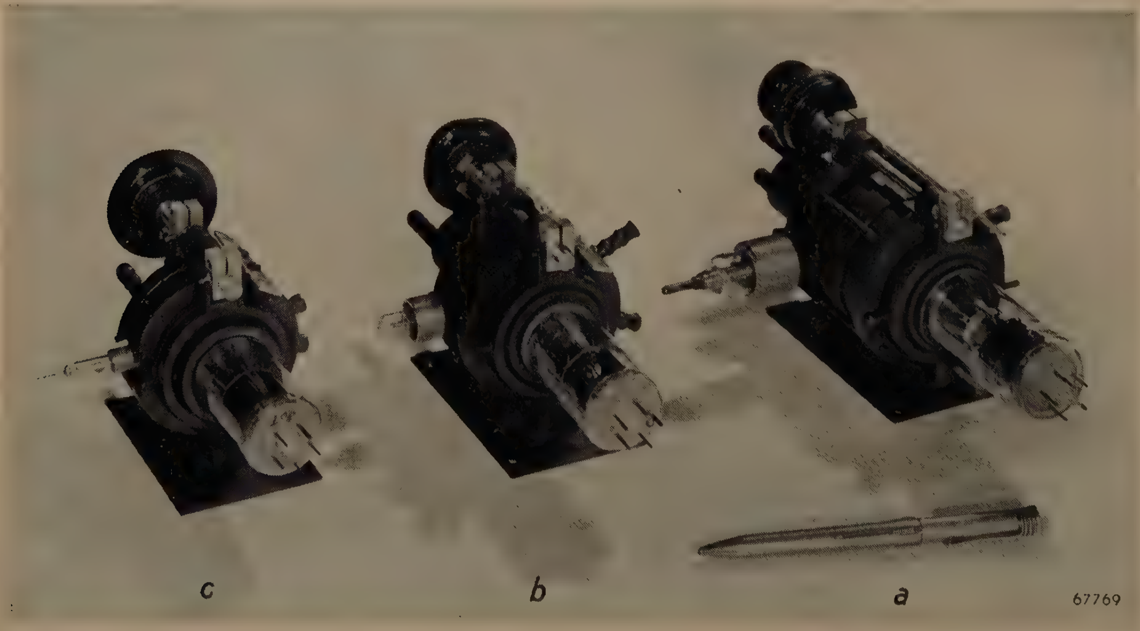

Fig. 17. Some velocity-modulation valves for different wavelength ranges: *a*) valve for $\lambda = 11$ to 15 cm, max. continuous output 500 W; *b*) valve for $\lambda = 5.5$ to 7 cm, max. continuous output 300 W; *c*) valve for $\lambda = 2.9$ to 3.6 cm, max. continuous output 100 W.

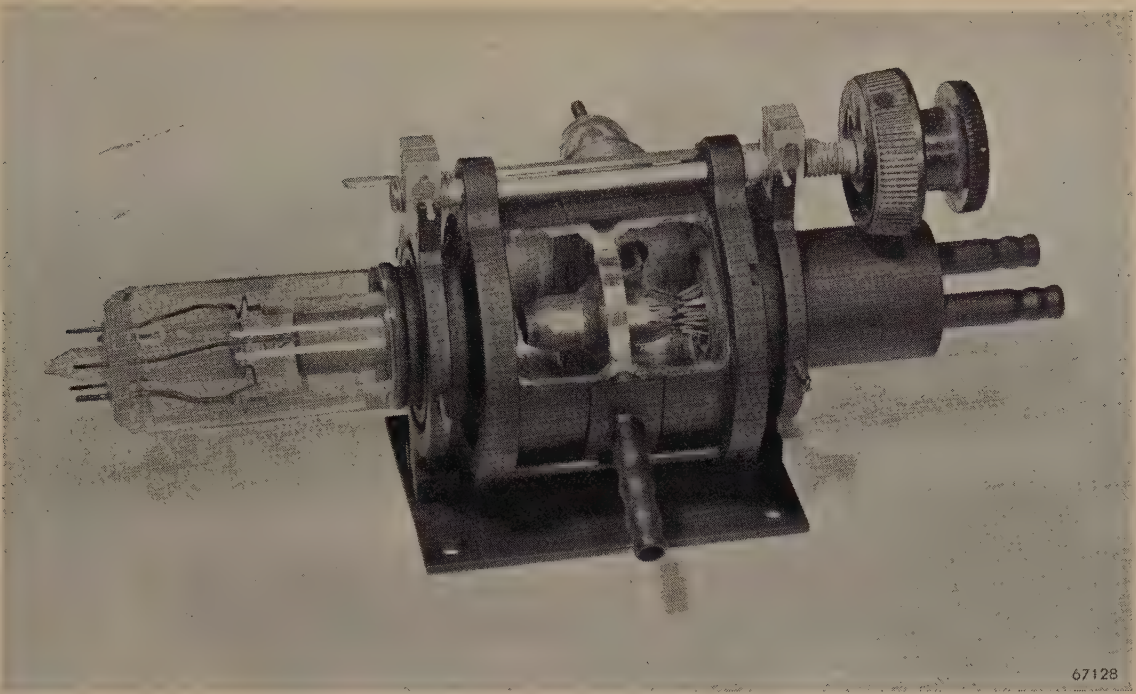

Fig. 18. Velocity-modulation valve for $\lambda = 11$ to 15 cm, with part of the wall of the cavity resonators removed, showing the toothed edges of the output gap.

175 mA, efficiency 20%). Here the inner conductor of the coaxial output system is not led out but is terminated inside the glass cap. This can be coupled into a rectangular wave-guide, as represented in *fig. 16*. Photographs of the complete valves are given in *figs. 17a* and *b*, while *fig. 18* shows a valve with part of the wall removed.

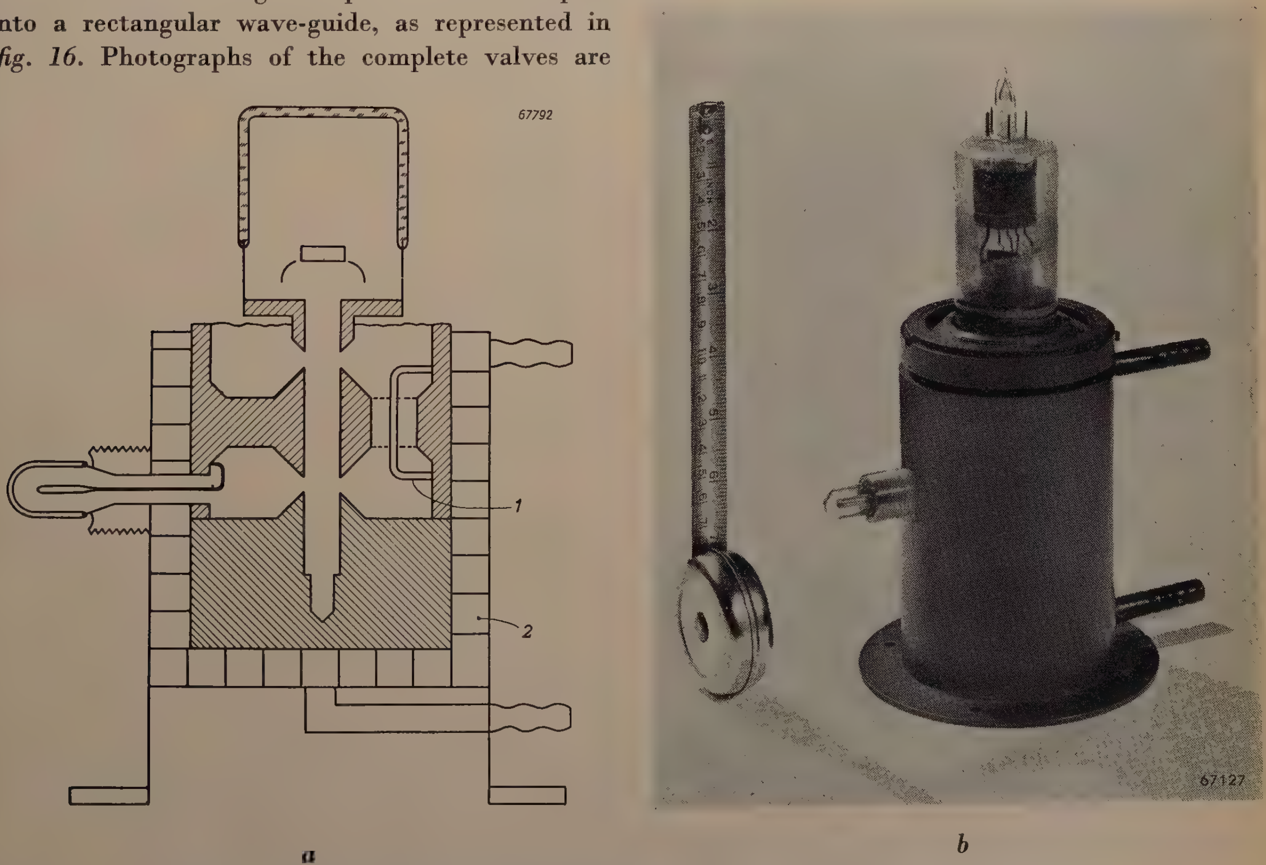

Fig. 19. Schematic cross section (*a*) and photograph (*b*) of a velocity-modulation valve for 10 cm waves and 1200 watts continuous output. *1* feedback bracket, *2* cooling jacket with helical partition.

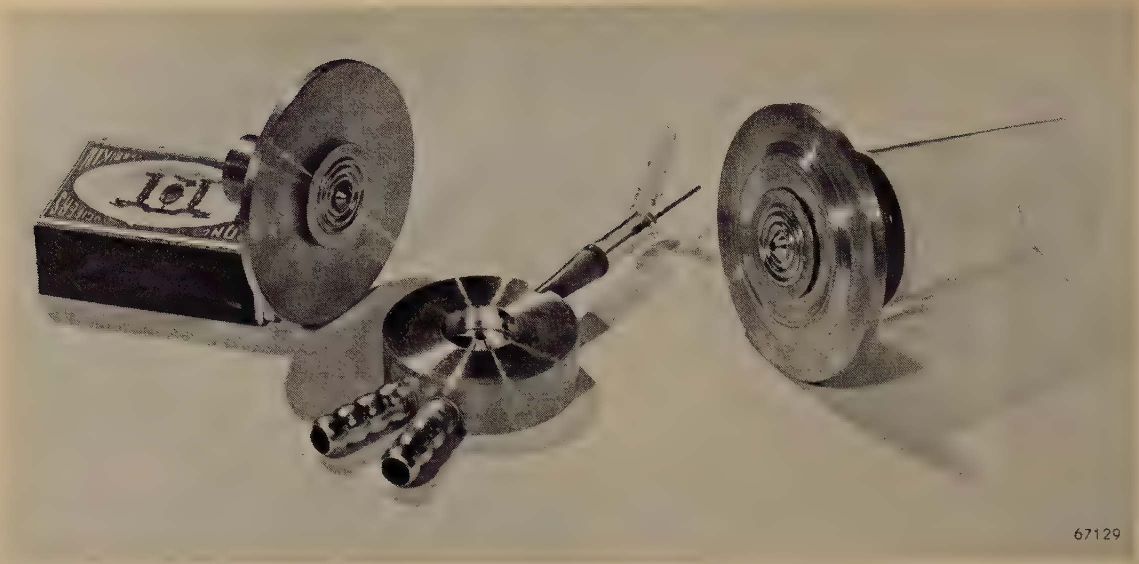

Fig. 20. The three parts making up a velocity-modulation valve for 3 cm waves. From right to left: cathode part with one of the flexible walls, middle part with output terminal and cooling jacket, anode part with the other flexible wall; the cooling jacket for the anode is put on later.

A third type of valve, without tuning, has been constructed for higher outputs than those mentioned above. The output cavity resonator has a fixed resonance, while the frequency of the input cavity can be varied only slightly for correct tuning. Further in this type the wall between the cavities has been made extremely thick (15 mm), with a view to improving the dissipation of the drift tube. As a consequence the mutual coupling of the cavity resonators cannot be effected by means of a hole alone — this hole would act as a wave-guide the cut-off wavelength of which would be much smaller than the wavelength at which the valve works. A bracket is therefore passed through the hole, forming a kind of coaxial feedback system (*fig. 19a*). The valve is enclosed over its entire length in a cooling jacket, to remove the heat produced by the H.F. losses in the cavity resonators — some hundreds of watts. On a wavelength of 10 cm, this valve, a photograph of which is reproduced in fig. 19*b*, yields an output of 1200 watts at a voltage of 15 kV and a beam current of 400 mA (efficiency 20%). Here the output system is similar to that used in the 6 cm valve, i.e. coaxial into a wave-guide (not passing through the glass envelope). In this valve the advantage of the L cathode is very striking, the filament current required being no more than 30 watts.

The fourth type of valve has been designed for wavelengths of about 3 cm. Its construction does not differ essentially from that of the other types. *Fig. 20* shows the three parts (cathode, middle, anode parts) of which the valve is made up, whilst in fig. 17*c* a photograph is given of the complete valve. This is suitable for wavelengths of 2.9 to 3.6 cm, yielding on 3.1 cm an output of 100 watts at a voltage of 6.6 kV, with a beam current of 150 mA (efficiency 10%). For constructional reasons the inner conductor of the coaxial output system is brought out, but here again the energy can be taken off via a rectangular wave-guide.

In this valve, too, the features of the L-cathode show to full advantage. The current density at the cathode surface under top load is 2.4 A/cm². With a former version of the L cathode used under these conditions a life of the order of 1000 hours was reached, and the latest version promises to have an even longer life.

Summary. Following a brief treatment of the elementary theory of the velocity-modulation valve, consideration is given to the problem of using the principle of velocity modulation for generating centimetric waves. It appears that in the construction of valves for large outputs the assumptions upon which the elementary theory is based are not sufficiently exact and that account has to be taken of some effects which had been disregarded in the theory. The adverse influence of one of these effects, the deflection of the electrons in the H.F. field of the input gap, can be reduced by giving the output gap a suitable shape. Further, by applying the L cathode, it is possible to simplify the construction of the electron gun. A description is also given of a new device for tuning the cavities. The article is concluded with a discussion of four experimental types of valves and the results obtained with them.

A THROUGH SUPERGROUP FILTER FOR CARRIER TELEPHONE SYSTEMS ON COAXIAL CABLE

by J. F. KLINKHAMER.

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Coaxial cables have been used in various countries for some considerable time in conjunction with carrier telephone equipment. Their importance in this field lies in the fact that hundreds of telephone channels can be superimposed on one pair of coaxial tubes. On the other hand, the provision of the necessary flexibility, in respect of the interconnection of blocks of channels at the nodal points of the cable network, has given rise to very severe technical problems. This difficulty arises partly from the multiplicity of channels, which was the raison d'être of the coaxial system, and partly from the requirements laid down by the C.C.I.F. at a rather early stage of development of coaxial technique. The through supergroup filter, described below, has been developed as a solution of some of these interconnection problems.

The transfer problem

During the past 15 years the long distance telephone plant of most countries has been undergoing a rapid expansion, in which carrier telephone systems have played an ever-increasing part. The growth of the carrier telephone network can be seen both in respect of the total route length of carrier cables and in the number of channels provided by each cable; although the number of conductors per cable has decreased, this has been more than compensated by the very large increase in the number of channels which can be provided on each pair of conductors. This figure has risen from 4, 12, 17 and 24 channel systems in the past [1]) to 48 and 60 channel systems on modern quadded carrier cable [2]) and to 600 or 960 channels on a coaxial system.

To discuss this latter trend of carrier technique is beyond the scope of this paper [3]). We are concerned here only with the transfer problems arising from the multiplicity of channels per pair, which occurs typically in the case of coaxial cable. *Fig. 1* shows as an example a line chart of a typical carrier system, in order to illustrate the above general statements. Suppose that between two main centres, c.q. the cities A and B, there is a carrier system capable of providing at least 1200 channels, of which the major part, say 720 channels, is required for communication between A and B, the remaining 480 channels being required for transit traffic, i.e. for communication from A via B to C and b_1 or from B via A to a_1. These transit channels must, therefore, be extended in station B to station C or b_1, and in station A to a_1.

If each of the A-B circuits were provided on the cable by means of a separate pair of conductors, then the circuits could be extended or terminated by simply connecting these conductors in any desired arrangement at A and B. No great difficulty is experienced if this method of connection is used on multipair carrier cables with, say, blocks of 48 [4]) or fewer channels per physical 4-wire circuit (one pair "go" and one pair "return"). Under these circumstances, each set of two pairs can be branched off or connected through in a relatively simple way, and complete blocks of channels, e.g. 48 channels, can be considered as units for traffic purposes.

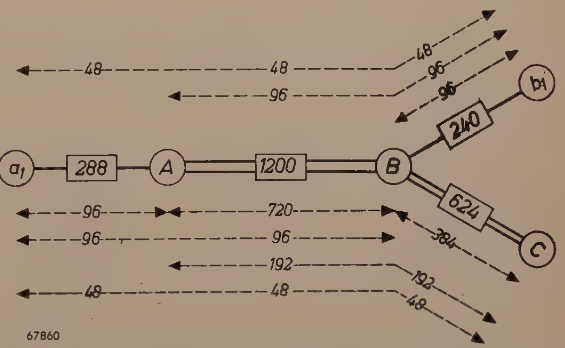

Fig. 1. Diagram of a trunk telephone network illustrating the problem of connecting circuits through on carrier cables. Between a_1 and A there are 288 channels available, between A and B 1200, etc. The numbers between the arrow points on the dotted lines indicate the distribution of channels to meet the traffic requirements.

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[1]) D. Goedhart and J. de Jong, Carrier-wave telephony, Philips Techn. Rev. **6**, 325-333, 1941.

[2]) G. H. Bast, D. Goedhart and J. F. Schouten, A 48-channel carrier telephone system, Parts I and II, Philips Techn. Rev. **9**, 161-170, 1947, and **10**, 353-362, 1948/1949.

[3]) An article on these subjects will be published in one of the forthcoming issues of this journal.

[4]) Telephone administrations usually round off the number of channels required to multiples of 12, 48 (as here) or 60, according to the transmission system under consideration.

In this way a reasonable approximation to the ideal arrangement is obtained, in which full use is made of all the available circuits on all sections of the network, as shown in *fig. 1*. When, however, coaxial cables are used, with say 600 channels on each pair of coaxial tubes, it is obvious that such a simple method can no longer be employed. Blocks of 600 channels are generally too large for approximating the traffic requirements between all switching centres.

Before proceeding to discuss the transfer problem on coaxial systems, the methods of modulation used with such system will be summarised, since this explanation is required in order that the technical problems associated with the transfer of small blocks of channels may be fully appreciated.

Modulation processes used on coaxial systems

The modulation processes used on coaxial systems are shown briefly in diagrammatic form in *figs. 2* and *3*. Fig. 2 is a simplified block diagram of the coaxial carrier equipment in a terminal station, whilst fig. 3 shows the various stages of modulation in the form of a frequency allocation diagram.

The subscriber is connected at point *I*. The speech frequencies lying in the band 0-4 kc/s are translated, by means of modulators and their associated filters, to the band of, say, 104-108 kc/s. This channel and 11 others — these latter lying in the bands 100-104, 96-100 kc/s etc. — are combined together at point *II* to form a group of 12 channels, which is thereafter treated as a unit.

The group is next transposed from the basic frequency range 60-108 kc/s [5]) by means of a second stage of modulation, say to the frequency band 312-360 kc/s. This group is then combined

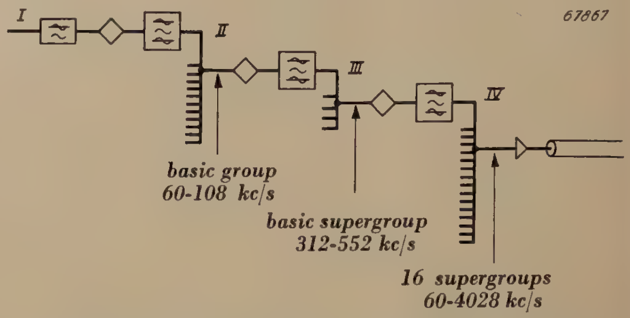

Fig. 2. Simplified block diagram of the modulation equipment in a coaxial carrier system. The subscriber is connected via the exchange at *I*. Twelve voice frequency channels modulate carriers and combine at *II* to form a group in basic position (60-108 kc/s). Five groups modulate five different carriers and combine together at *III* to form a supergroup in basic position (312-552 kc/s). Ten or sixteen (future developments may increase this number) supergroups each modulate different carriers and combine at *IV*. This combination of supergroups is then transmitted via the outgoing coaxial tube. ⊟ low-pass filter, ⊟ band-pass filter, ◇ modulator, and ▷ amplifier.

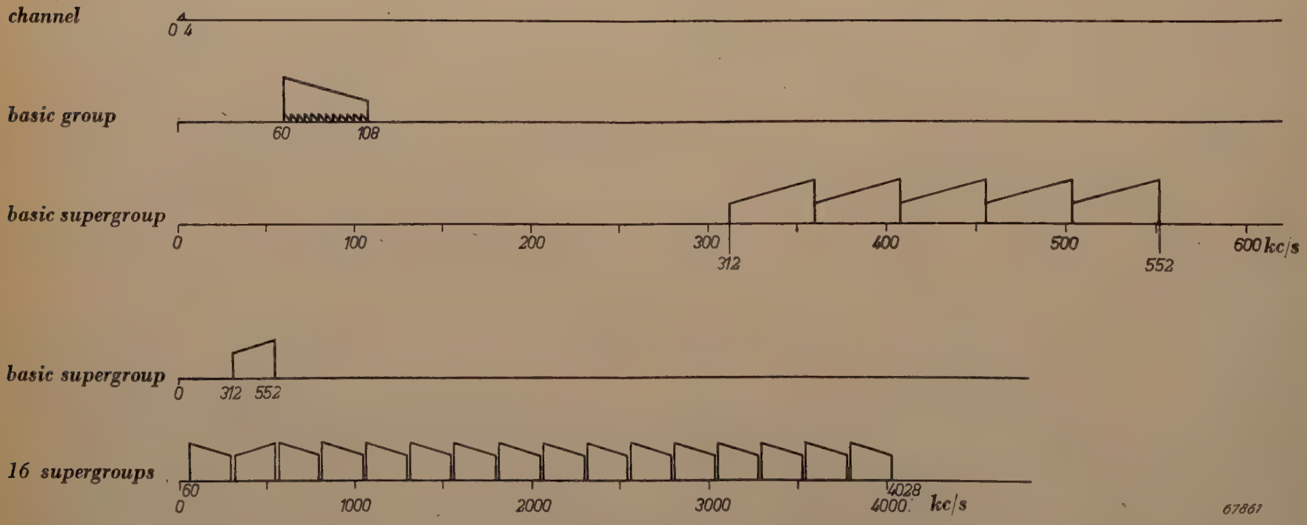

Fig. 3. Frequency allocation diagram showing the modulation processes of fig. 2. The first three stages show the positions occupied in the frequency spectrum by a voice frequency channel, a group of twelve channels in basic positions and a supergroup of five groups in basic position. The last two lines show, on a reduced frequency scale, the supergroup in basic position, and the combination of 16 supergroups sent to line. The ascending lines indicate the direction of increasing voice frequency in the channels, so that it can be seen for each case whether the upper or lower sideband has been chosen.

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[5]) The position of the basic groups has been fixed by international agreement, in accordance with the recommendation of the Comité Consultatif International de Téléphonie (C.C.I.F.) that it should occupy either the band 60-108 kc/s or the band 12-60 kc/s, for international, and if practicable, for national traffic. The system described in this article uses a basic group occupying the 60-108 kc/s position, but all the considerations given here apply mutatis mutandis to a system using the other basic group. The division into groups in general, and also the division into supergroups mentioned later in the text, are both based upon C.C.I.F. requirements.

at *III* with four other groups, which have been translated to the bands 360-408, 408-456 kc/s etc. This combination of five groups is called a supergroup and is also subsequently treated as a unit.

The supergroup is translated from its basic frequency range (312-552 kc/s), by means of a third stage of modulation, to the frequency range it must ultimately occupy on the coaxial line. At *IV* several supergroups are combined together, say 10 or 16, and, after amplification, are transmitted to the coaxial cable. The final positions occupied by the supergroups on the line are such that there is a space of, in most cases, 8 kc/s — the width of two channels — between adjacent supergroups. The reason for this space will be explained later in the text.

The through group filter

Let us now return to the transfer problem: taking a relatively simple case of the general type shown in fig. 1, assume that there are only 12 transit circuits to be extended from a cable *AB* to a cable *BC*.

It can be seen from this diagram that the 12 channels (which form a complete group on both cables *AB* an *BC*) occupy the same frequency range at both points *III* and *IV*. It might be asked whether the points *III* and *IV* could be directly connected together without the modulation equipment shown.

The impracticability of this may be seen from *fig. 5*. The group filter (GF_4 in fig. 4), which selects the transit group, in this case group 4, from the supergroup concerned, has an attenuation/frequency characteristic as shown in fig. 5*b*. The filter not only passes group 4 unattenuated, but also passes parts of the adjacent groups 3 and 5 with varying degrees of attenuation. If the points *III* and *IV* were directly connected together, then those parts of the adjacent groups 3 and 5 passed by the group filter would leave section AB and reach the junction of *III* and *IV*, on either side of the basic frequency range. After remodulation of the group containing the transit channels, say to the position of group 3′ in another supergroup, for transmission to *C*, the vestiges of the previously adjacent groups are further attenuated by the group filter GF_3, but are still at a relatively high level at the point *V*, where they will fall within the frequency bands of groups 2′ and 4′ of the supergroup formed at *B* (see figs. 5*g* and *h*). Conversations in a number of *A*-*B* channels may therefore be overheard on *B*-*C* channels, at such a level as to be intolerable.

The above effect can be attributed to the shape of the filter characteristic in the transition band,

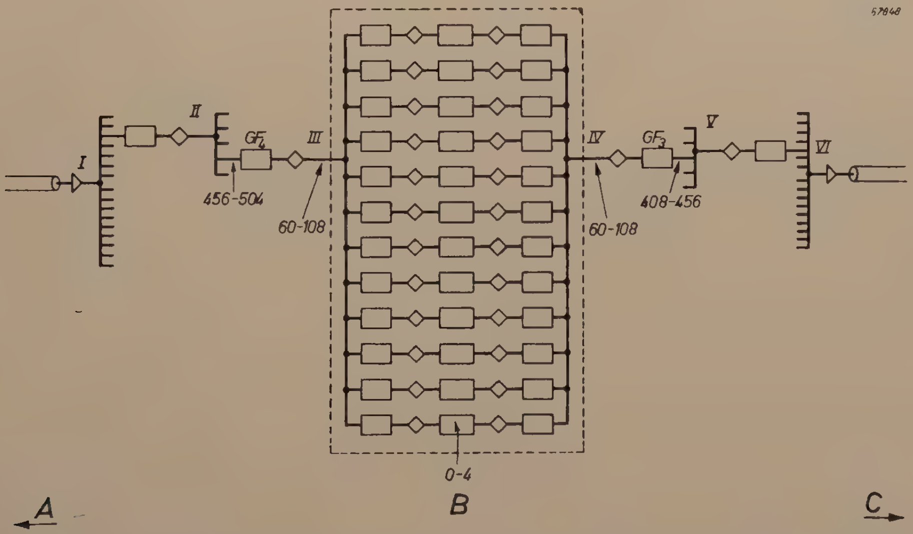

Fig. 4. Schematic block diagram showing how the fourth group of the third supergroup from station *A* is connected through at station *B* to station *C* as the third group of the fifth supergroup, when all the other groups terminate at *B*. If the "through group" is completely demodulated and remodulated, all the equipment between *II* and *V* is required.

One obvious solution is for these 12 incoming channels, as well as all other channels terminating in *B*, to be demodulated there into separate audio frequency channels; the transit channels may then be remodulated for transmission to *C*. At *B*, therefore, there must be two complete sets of terminal equipment, arranged back to back. *Fig. 4* shows, in block schematic form, the equipment required for the 12 through circuits.

i.e. the region between the pass band and the stop band. If these filters had a characteristic more nearly resembling that of an ideal filter (small constant attenuation over the pass range, rising steeply to a high value of attenuation in the attenuating range) then points *III* and *IV* could be connected direct together and the intermediate apparatus (which ensures that the transit group is free from vestiges of other groups) could be omitted.

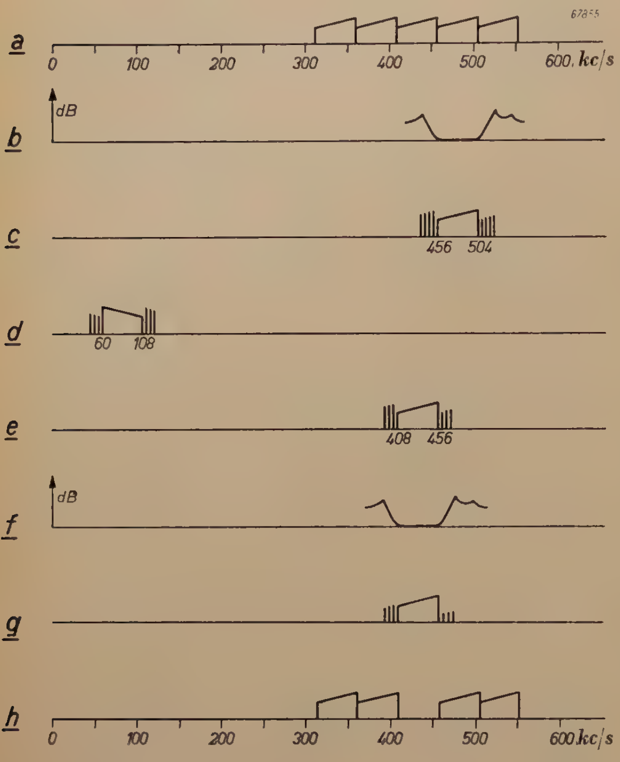

Fig. 5. This diagram illustrates the impracticability of connecting points *III* and *IV* of fig. 4 directly together. *a*) The supergroup at point *II*. *b*) the attenuation/frequency characteristic of the group filter GF_4. *c*) Group 4 after traversing the group filter GE_4. The vertical lines on each side of this group denote vestiges of the adjacent groups (3 and 5) which pass through GF_4. *d*) The through group demodulated to the basic position at point *III* of fig. 4 and, if both are connected, also at point *IV*. *e*) The group at the input of the group filter GF_3. *f*) Attenuation/frequency characteristic of the group filter GF_3. *g*) The group, transferred as group 3, at point *V*, is combined with four other outgoing groups (*h*), to form a supergroup.

As will be seen later, it is not a practical proposition to make the characteristics of all the group filters conform to such a high standard of performance. One of the principal advantages resulting from the use of a multiple modulation system is, in fact, the relative simplicity of the group filters.

It is possible, however, in the case of a transit group, to supplement the effect of the two group filters by means of a special filter, having a characteristic more closely approximating to the ideal. If this latter filter has a pass range of 60-108 kc/s, it can be connected direct between points *III* and *IV*, in place of the equipment represented by that part of fig. 4 enclosed within the dotted line. Since every group required for transit connection would be demodulated to occupy the basic frequency range 60-108 kc/s, only one type of this filter would be required.

Such "through group filters" have been constructed [6]). Although it is impracticable to construct an ideal filter, it is possible to make a filter which approximates to the ideal; but the difficulties increase in proportion to the mid-band frequency at which the filter must operate, the ratio of the pass band to the transition band and the dissipation factor of the coils and capacitors. In the case in question the requirements are rather severe. The pass band is 48 kc/s wide, whereas the transition range between the pass and attenuating regions is only 0.7 kc/s. This figure of 0.7 kc/s is the same as that for the channel filters, which only pass a band nominally 4 kc/s wide and the majority of which work at relatively low frequencies. The attenuation required in the stop bands of the 48 kc/s wide filter is at least 70 db.

It would not be practicable, with components or techniques available at present, to construct a group filter answering these requirements for operation in any of the five frequency bands 312-360, 360-408, 408-456, 456-504, or 504-552 kc/s. Filters having a pass band corresponding to the basic group of 60-108 kc/s, with steep transition characteristics, can, however, be made. For this purpose filters with quartz crystals are employed. These filters will be dealt with presently, but it will be useful first to extend the line of thought developed.

Through supergroup filter

When it is desired to transfer at station *B* more than one group, say two or three groups, two or three through group filters, respectively, are used. When however it is desired to connect through a complete supergroup, the possibility of omitting some of the modulation equipment has again to be considered. From *fig. 6* it can be seen that at points *II* and *V* the supergroups all occupy the basic supergroup band of 312-552 kc/s. As in the case of the group filter already mentioned, the supergroup filter *SF* will pass portions of the adjacent supergroups with little attenuation, preventing thereby the point *II* from being connected direct to point *V*. There is, however, as before, the possibility, at least in theory, of supplementing the normal supergroup filter *SF* by means of a special "through supergroup-filter". Such a filter would replace all the apparatus contained within the dotted line on the block diagram of fig. 6. The action of such a filter is shown in the form of a frequency allocation chart in *fig. 7*.

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[6]) C. F. Floyd, Journal of the Institution of Post Office Electrical Engineers **15**, 57-62, 1947, (July); E. F. S. Clarke and T. F. Reed, ibidem **15**, 109-114, 1947 (October).

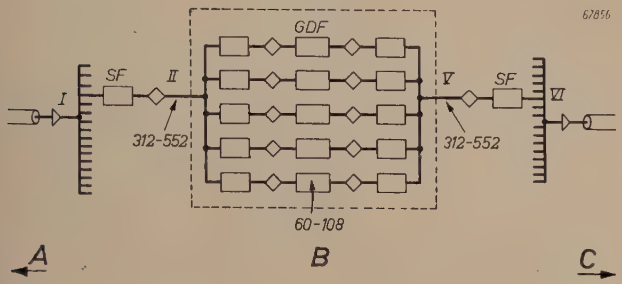

Fig. 6. A complete supergroup (No. 4) connected by means of five through group filters *GDF*, supplementing the action of the supergroup filter *SF*. One single filter, however, in place of the equipment within the dotted line, can serve the same purpose.

The design of a through supergroup filter, fulfilling all the necessary requirements, is a difficult task. It is true that a much wider gap is available between adjacent supergroups for the transition bands between the pass and attenuating regions, than in the case of the through group filter. (This gap, which is 12 kc/s on either side of the supergroup in the basic position and 8 kc/s between all other supergroups, was in fact left as a provision for the use of through supergroup filters.) Offsetting this advantage, however, is the fact that the filter must be capable of passing a band 240 kc/s wide as well as working at the relatively high frequencies of 312-552 kc/s.

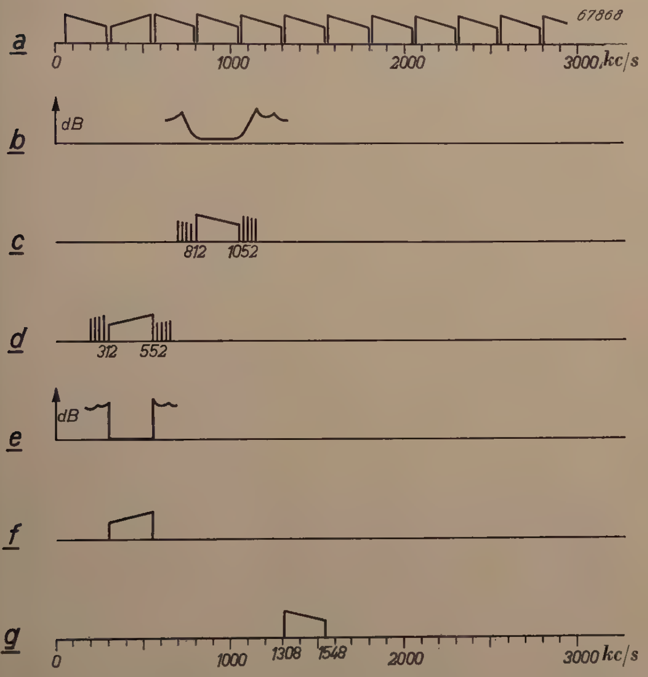

Fig. 7. A supergroup (No. 4) connected through by means of a through supergroup filter. *a*) Incoming supergroups at point *I* of fig. 6. *b*) Attenuation/frequency characteristic of the supergroup filter *SF* for the fourth supergroup. *c*) The "through" supergroup extracted by *SF* with vestiges of the adjacent supergroups. *d*) The through supergroup modulated to the basic position (point *II* in fig. 6). *e*) The attenuation/frequency characteristic required for a through supergroup filter. *f*) The through supergroup at point *V* with the vestiges of adjacent supergroups removed. *g*) The through supergroup, transposed to its new position on the outgoing coaxial cable (point *VI*).

Moreover, through supergroup filters have to fulfil other requirements which can best be summarised from a British Post Office specification.

The insertion loss, i.e. the attenuation caused by the insertion of a filter between a voltage source and a consumer, in the pass band may not vary, as a function of frequency, by more than ± 0.25 dB, over the whole of the pass band; within the frequency range of each group (48 kc/s wide) this variation may not even exceed ± 0.125 dB. The discrimination (i.e. the difference between the attenuation in the pass and suppression regions) has to be at least equal to the values shown in *fig. 8*. At 308 kc/s and 556 kc/s in the transition regions, the discrimination must be at least 40 dB. When two sinusoidal signals, having frequencies lying in the pass band, are applied at levels of 30 dB below a milliwatt to the input of the filter, the filter being terminated at both input and output in 75 ohms, then the intermodulation products due to non-linearity of the filter elements must be at least 75 dB below the output level of either of the two signals.

The underlying reasons for these British Post Office requirements are as follows.

The high discrimination at 308 kc/s and 556 kc/s is required because line pilots are transmitted in the gaps between some supergroups for purposes of test and regulation[7]. Such line pilots — sinusoidal signals whose input level and frequency are accurately maintained — are transmitted over the whole distance from the sending to receiving terminal stations, i.e. from the modulating equipment at the sending terminal to the equipment in the receiving terminal where the supergroups are separated by demodulation. After demodulation of a supergroup any adjacent line pilot will appear as one of the frequencies 308 kc/s or 556 kc/s. When a supergroup having an adjacent line pilot is transferred by the through supergroup filter method, these frequencies of 308 kc/s or 556 kc/s might coincide, after remodulation, with one of the line pilots used on the outgoing cable. Consequently, the incoming pilot after demodulation has to be sufficiently attenuated in the through supergroup filter to avoid any such leakage which might ensue and cause an intolerable error in, or the possibility of beats with the outgoing pilot.

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[7] This is another reason for leaving an 8 kc/s gap between adjacent supergroups.

The requirements with respect to the attenuation variation in the pass band are of greater importance for the present consideration. The reasons for these requirements are based on considerations somewhat different from those already discussed in connection with the cable network of fig. 1.

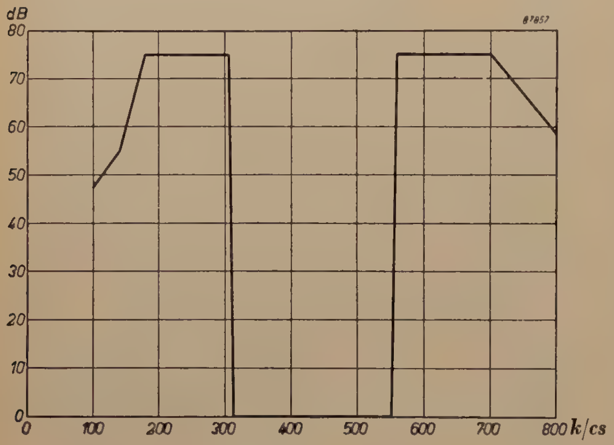

Fig. 8. Minimum discrimination to be given by the through supergroup filter outside the band-pass range (312-552 kc/s).

In this network an example was given in which the majority of the circuits were carrying traffic from *B* to *A* (720) and from *B* to *C* (384), whilst a relatively small number of transit circuits were connected through from *A* to *C* (192). Now let us consider an example of the opposite nature. Suppose that in fig. 1, instead of the numbers of channels stated, the majority of the traffic is between *A* and *C*, and that only a few channels, say about 60, are required for traffic from *B* to *A*, and a similar number for *B* to *C*. *B* is thus considered as a minor intermediate station on the route *AC*. Under these circumstances, it would be preferable to connect the cable *AB* to the cable *BC* direct, and to extract at *B* the one supergroup required (*fig. 9*). In order that the same frequency band can be used for this supergroup in both sections of the route, a band suppression filter, having a stop band in the relevant frequency range, must be connected in the position shown by the dotted block in fig. 9 [8]). Such a filter would allow the frequency bands used for transit channels to pass with little or no attenuation. The C.C.I.F. requirements in respect of supergroup spacing are, however, such that a filter of this type cannot be made for the majority of the supergroups [9]).

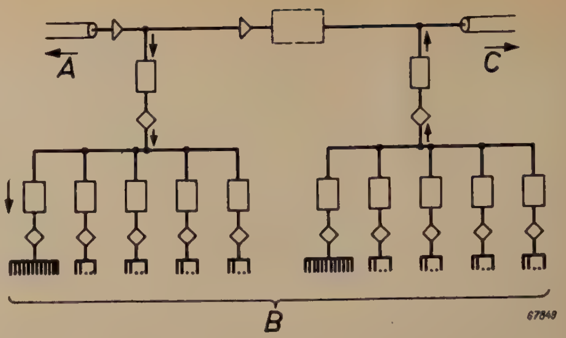

Fig. 9. When at station *B* all channels except one supergroup have to be connected through, it would be advantageous to join the incoming and outgoing coaxial cables, via a supergroup stop filter (shown dotted). It is not yet feasible, however, to make such a filter for other supergroups than Nos. 1 and 2.

One not altogether attractive but practicable solution to this problem is to demodulate all the supergroups at *B*, so that they occupy the basic supergroup frequency band of 312-552 kc/s. All the supergroups carrying transit traffic, i.e. all except the *A-B* supergroup to be extracted, are passed through separate through supergroup filters, and then remodulated to their former positions in the frequency spectrum. The same frequency band is used for the *A-B* and *B-C* supergroups; it is obvious that circuits are not connected via a through supergroup filter but terminate at *B* in two separate sets of terminal equipment (see *fig. 10*).

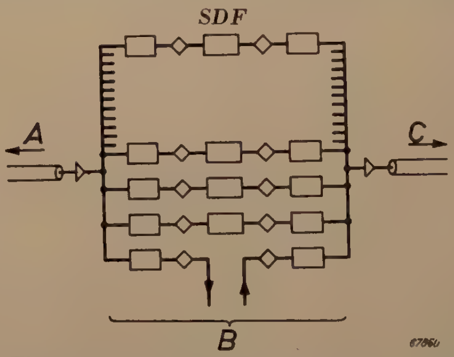

Fig. 10. The same as fig. 9, but with all the through supergroups routed via through supergroup filters *SDF*.

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[8]) There are some other possibilities, such as the practice of "branching" or "echelon working" (to borrow a term from Baudot telegraphy) in which the frequency range of the channels that are extracted at an intermediate point is not used again for any other channel for the remaining part of the coaxial tube. Such solutions will, however, be left out of consideration here, as stated in the beginning of the article.

[9]) One might suggest to translate the frequency band of the complete set of supergroups on the coaxial tube bodily, so that a given supergroup may be brought into a frequency range such that a band stop filter may be constructed. For technical reasons, however, which are irrelevant to this paper, this is not practicable.

Thus at a station where some of the channels have to be extracted, the majority of the circuits pass through the station via supergroup modulators and through supergroup filters. On coaxial backbone routes, therefore, on which there are a number of minor intermediate stations or small branches at each of which a few circuits have to be terminated, the important long-distance circuits have to pass through a succession of through supergroup filters.

A multiplicity of through supergroup filters and supergroup translation equipment will also occur in coaxial systems which are arranged to give alternative routing, by means of a triangulated network of cables, between towns of fairly uniform size.

From this explanation it will be seen that the severe requirements for through supergroup filters arise partly from their use in cases where they are not the logical solution of the problem, but where they have to be used because no other solution is available.

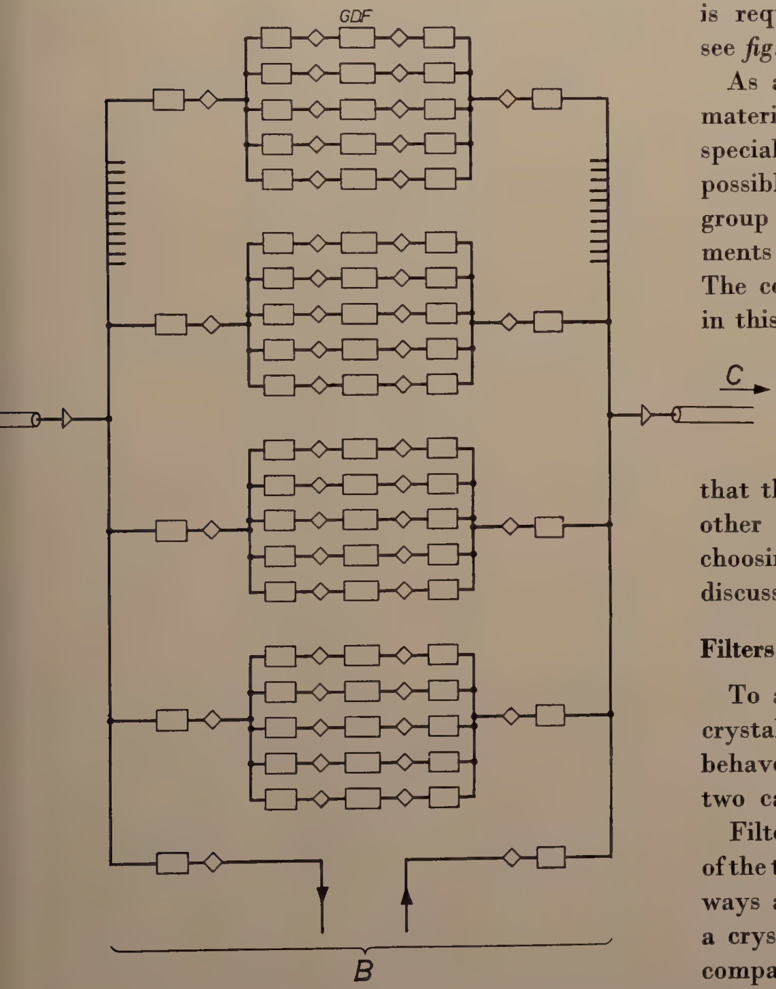

Fig. 11. The method which had to be employed, using through group filters (*GDF*), before a through supergroup filter was available. This method requires about five times the equipment necessary with the through supergroup filter method shown in fig. 10.

Because variations of attenuation in the pass bands of a number of similar filters in tandem are additive, it will readily be understood why the variations of attenuation in the pass range of the through supergroup filter must be kept within such narrow limits. Intermodulation in the filters is also cumulative (though in a different way), which further justifies the corresponding requirements of the specification.

As far as is known no through supergroup filter has hitherto been made which satisfies the above specification in respect of this variation of attenuation in the pass band. When a supergroup had to be extracted at an intermediate point, the only solution available in the past, if waste of frequency band (and revenue) were not permissible, was to demodulate the entire traffic on the coaxial cable down to basic groups. The C.C.I.F. contemplated the necessity of this operation at several points along the length of a typical long distance circuit. This method requires about five times the apparatus that is required for the through supergroup method, see *fig. 11.*

As a result of the use of Ferroxcube as a core material for filter coils, and of the application of special methods of computation, it has now become possible to design and construct a through supergroup filter which completely fulfils the requirements of the above mentioned B.P.O. specification. The construction of such a filter will be discussed in this paper in general terms.

In order to fulfil the requirement that a through supergroup filter must have a characteristic approximating as closely as possible to the ideal, it might be thought that the best solution would be to use quartz or other piezo-electric crystals. The reasons for choosing an alternative solution will now be discussed.

Filters with quartz crystals

To a first approximation a piezo-electric quartz crystal placed between two electrodes, *P* and *Q*, behaves electrically as a combination of a coil and two capacitors connected as shown in *fig. 12.*

Filter networks, having sections containing units of the type shown in fig. 12, can be designed in various ways and every such unit can then be realised by a crystal. Such crystal filters have the advantage, compared with those using coils and capacitors, that the elements in the equivalent circuit of the crystal have only small loss angles (*Q* of between 1000 and 300,000), so that a close approximation to the ideal characteristic can be obtained. Furthermore, if the crystals are properly ground, the equivalent inductance and capacitances are practically independent of changes of temperature. Because of these advantages there is an important field of application for crystals in the manufacture of filters, despite their higher cost and more fragile nature compared with the more usual coil and capacitor combination.

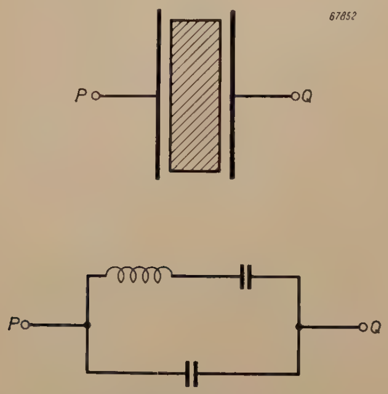

Fig. 12. A crystal clamped between two electrodes behaves electrically as a combination of an inductance with two capacitances as shown above.

Several types of through group filters have been constructed using quartz crystals. In these filters, as is generally the case with filters required to have a relatively wide pass band, a difficulty peculiar to the use of quartz crystals has to be overcome. A crystal slab can vibrate in more than one mode and it is not always possible to suppress undesired modes. If such modes occur, the simple circuit of fig. 12 no longer holds, and an equivalent circuit composed of many more elements is in fact obtained. Each undesired mode of vibration within the pass band gives rise to a narrow, sharp peak of attenuation at a particular frequency. It is extremely difficult to eliminate these peaks from the pass band when crystals are used in wide-band filters.

Some of the through group filters at present in use have this disadvantage, which, however, has been partially overcome by arranging that the undesired peaks of attenuation lie centrally in the space between adjacent channels of the group. Between two adjacent channels of a group there is an unused frequency band of 700 c/s, and the unwanted peaks of attenuation, being narrower than 600 c/s, do not cause any interference.

This artifice cannot, however, be used in through supergroup filters, because the prescribed pass band is much wider and at a higher frequency; the undesired attenuation peaks of the crystal under these circumstances are also much wider, with the result that some of the channels in a supergroup could not be used. The gaps which would have to be left, to avoid the resonance peaks, would conflict with the purpose for which the through supergroup filter is intended, namely to preserve the full flexibility of the system, without loss of channels and the practical complications arising therefrom.

Design of the through supergroup filter

From the above it is seen that until a few years ago it was extremely difficult to meet the requirements for a through supergroup filter, because of the undesired resonances when using crystals, and the excessive losses when using coils and capacitors. The use of the new magnetic ceramic material Ferroxcube for the cores of filter coils, however, has now provided a very satisfactory solution to the problem of building a coil-and-capacitor through supergroup filter. Coils having a Q of 350 at 550 kc/s can easily be manufactured using this core material. The best Q obtainable with coils of comparable dimensions in the past was 200. The loss angle of normal mica capacitors is small enough for use in through supergroup filters.

It was necessary to use the frequency parameter method, rather than the more usual image impedance or stencil methods, in designing the filter. A detailed discussion of these methods will not be attempted in this paper [10]), but it should

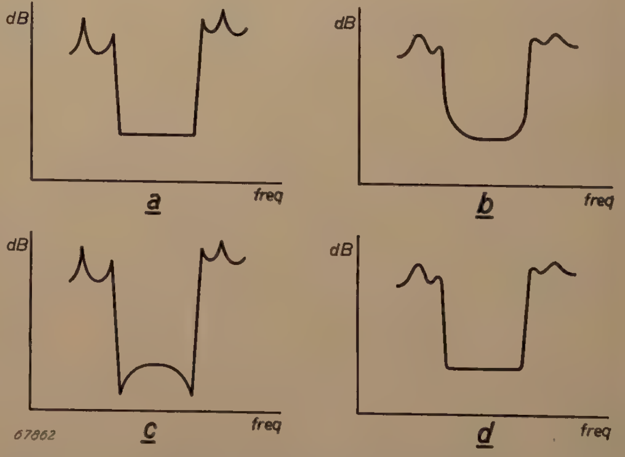

Fig. 13 *a*) Attenuation/frequency characteristic of a filter, designed by the image impedance method, with ideal, loss-free, coils and capacitors, and terminated by the image impedance. *b*) As (*a*) showing the effect of using practical components and terminating with a constant resistance. *c*) Attenuation/frequency characteristic of a filter designed by the frequency parameter method showing over-equalisation obtained with loss-free components. *d*) As (*c*) showing the effect of using practical components. Note the sharp corners at the edges of the pass-band range.

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[10]) An explanation of these methods with particular reference to the calculations for the through supergroup filter dealt with in this article, will appear in Philips Research Reports.

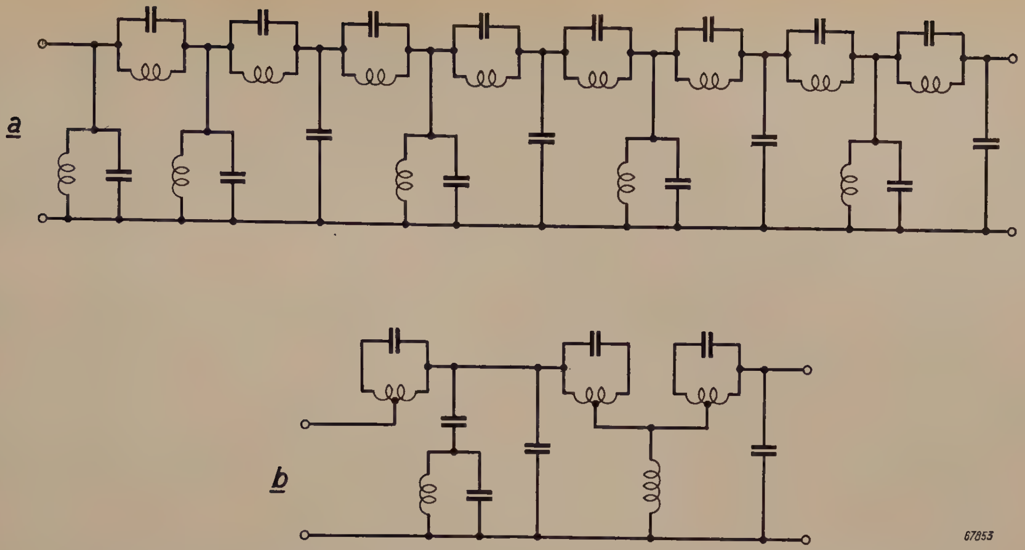

Fig. 14. *a*) Circuit diagram of half of the through supergroup filter. *b*) Circuit diagram of the supplementary pilot stop filter.

be noted that the latter two methods result in a filter in which there are no abrupt changes of attenuation, i.e. the change of attenuation from the pass band to the transition region is represented as a smooth curve on the attenuation/frequency characteristic, and covers a relatively wide frequency range. A filter can be designed, using the image impedance or stencil methods, having sharp changes of attenuation, but only at the cost of using a large number of coils and capacitors (see *fig. 13a* and *b*). It is possible, however, to design a filter, using the frequency parameter method, and assuming loss-free components, having a characteristic in which there is over-equalisation near the edges of the band (fig. 13*c*). It can then be arranged to use this excess equalisation so that with components having predetermined loss angles, the pass-band characteristic becomes flat, with sharp corners (fig. 13*d*).

In the course of the filter calculations for this specific case, the attenuation requirements were distributed over three correlated filter networks, as follows.

Disregarding initially the required discrimination at the pilot frequency of 556 kc/s, a through supergroup filter was designed having the desired pass band of 312-552 kc/s. The variation of the attenuation over the whole of the pass band of this filter was not greater than ± 0.125 dB, whereas the discrimination between the pass and attenuation bands was 40 dB. A schematic representation of this filter is shown in *fig. 14a* and its measured attenuation/frequency characteristic in *fig. 15*.

Two of these networks in cascade, separated

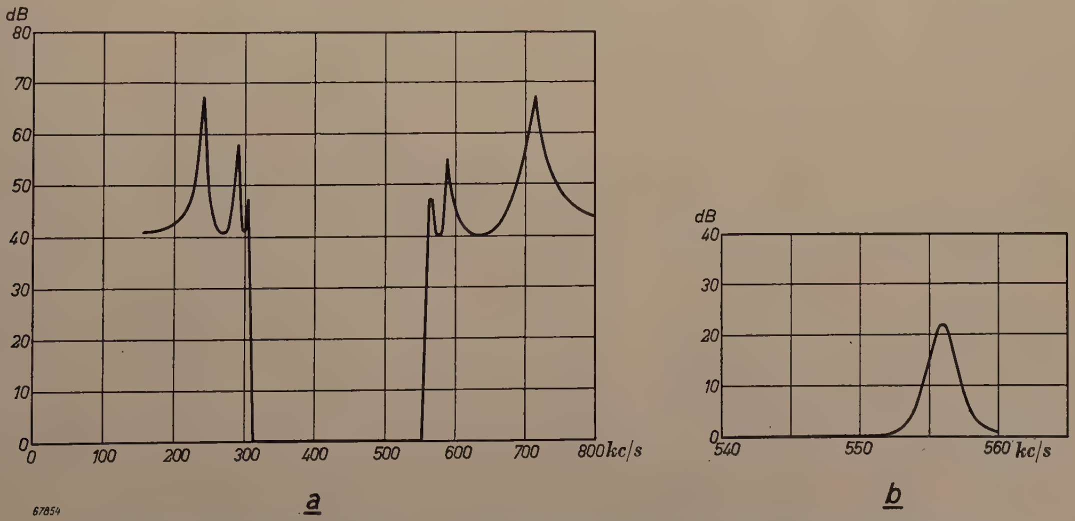

Fig. 15 *a*) Attenuation/frequency characteristic of half of the through supergroup filter. In this figure and in fig. 16 the attenuation in the pass band has been taken equal to 0 *dB*, see also fig. 20. *b*) Attenuation/frequency characteristic of the pilot stop filter.

either by an amplifier [11]) or by a resistance pad, in order to avoid interaction loss caused by the somewhat irregular impedance of the filters, fulfil all the requirements for a through supergroup filter except for the discrimination in the vicinity of 556 kc/s, which is about 20 dB too small. This requirement is met by an additional pilot stop filter which has a constant attenuation in the 312-552 kc/s range, but has a peak of attenuation of about 20 dB at 556 kc/s. The circuit diagram and attenuation/frequency characteristic of this filter are shown in figs. 14*b* and 15*b* respectively. The fully drawn curve of *fig. 16* shows the attenuation/frequency characteristic of the complete filter, whilst the dotted curve on the same figure shows the minimum specification requirements (see fig. 8), for purposes of comparison.

During further development it became apparent at an early stage that a choice had to be made for practical reasons between sacrificing about 10 dB of the discrimination in the narrow range between 560 kc/s and 561 kc/s or introducing a further stage of amplification (or alternatively a special transformer). For the filter described in this paper the first solution was chosen, with the result that in the 560-561 kc/s band the requirements of the specification are not fully satisfied [12]). In the rest of the attenuation range however, the requirements of the specification are everywhere exceeded by at least 5 dB.

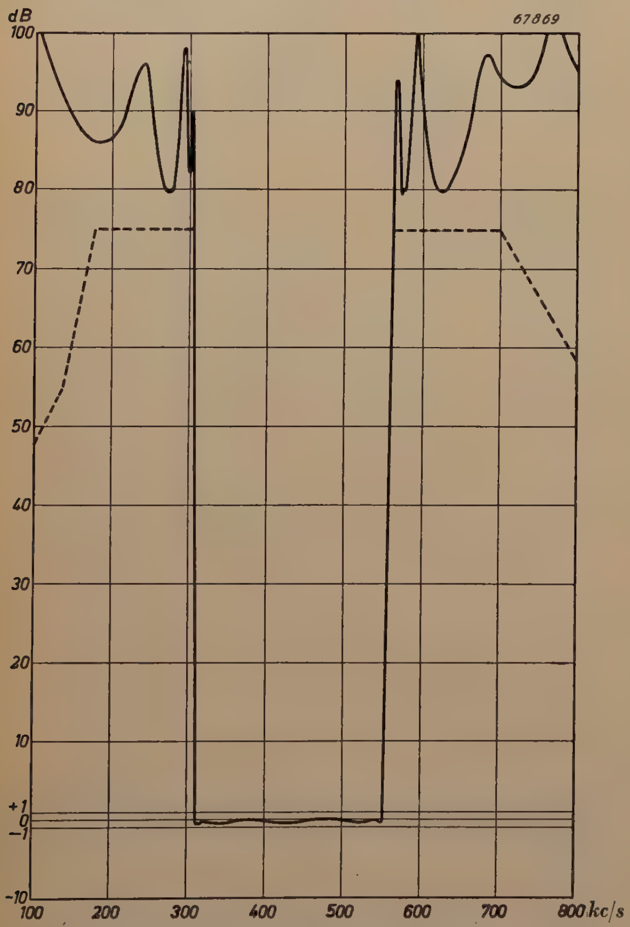

Fig. 16. Attenuation/frequency characteristic of the complete through supergroup filter. The dotted lines indicate the required minimum discrimination, see fig. 8. The specified minimum attenuation in the stop-bands is exceeded by 5 dB or more at all frequencies except in a very narrow frequency range (hardly perceptible on this diagram, because of the scale), from 560-561 kc/s.

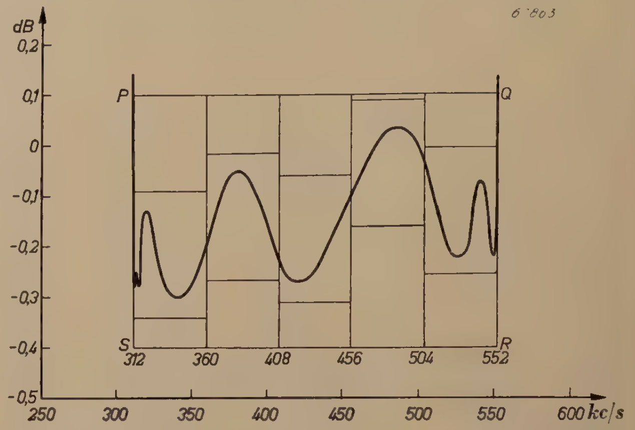

Fig. 17. Attenuation of the through supergroup filter in the pass band. The attenuation scale has been enlarged, so that the variations in attenuation can be seen clearly. The diagram shows that the attenuation variation as a function of frequency remains well within the specified limits, both for the whole supergroup (rectangle *PQRS*) and for each of the groups separately (5 smaller rectangles.)

In order to show how the requirements for the pass-band range have been met, this part of the attenuation/frequency characteristic has been reproduced in *fig. 17* with a greatly enlarged attenuation scale. From this figure it can be seen that the total variation of attenuation in the pass band does not exceed 0.5 dB, since the curve lies wholly within the rectangle *PQRS*. The filter thus fulfils the pass-band requirements. It can also be seen from this figure that the total attenuation variation in each of the five groups (the small rectangles) does not exceed 0.25 dB and thus the requirement that the attenuation should not vary by more than ± 0.125 dB in any one group is satisfied.

In addition, it may be mentioned at this point that the intermodulation products are all at least 80 dB below the output level of the original signals when these latter are applied at an input level of 30 dB below 1 mW.

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[11]) By means of this and other amplifier stages incorporated in the filter, it is also possible to arrange for the "filter" to have a gain (negative pass-band attenuation).

[12]) As a consequence, signals of audio frequencies 0-1000 c/s of the channel 560-564 kc/s of the adjacent supergroup will be able to penetrate without sufficient attenuation into a corresponding channel in the outgoing cable. This limited crosstalk, which is worst in the vicinity of 500 c/s, is not very important. If considered necessary, the other alternative might still be adopted.

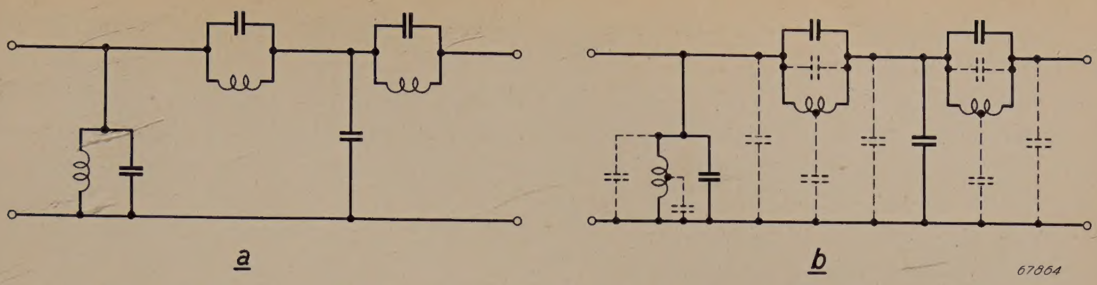

Fig. 18 *a*) Schematic diagram of a typical section of a filter. *b*) The same as (*a*) but including the stray capacitances.

Owing to the closeness with which the impedances of the network had to approximate to the calculated values, and in addition the relatively high working frequencies, the stray capacitances of the wiring, windings etc., had to be taken into account in the design. *Figs. 18a* and *b* show a typical network with and without the most important stray capacitances. The stray capacitances shown are (1) self capacitances of the coils, (2) capacitance between coils or capacitors and earth, and (3) capacitance between wiring junctions and earth.

The way in which all such stray capacitances have been reduced to values which can be tolerated, is in principle quite simple. From the number of possible alternative configurations in which the various circuit units in a filter can be built up, a circuit design has been chosen in which all stray capacitances appear across theoretical capacitances. The value of the actual mica capacitor is then so chosen that the combined capacitance is equal to the theoretical design value. The design of the filter must obviously be such that the theoretical capacitances required are in all cases higher than the stray capacitances. In the three filters mentioned above, therefore, it is necessary to have a capacitor in parallel with each inductance (case 1) and in addition a capacitor between each wiring junction and earth (case 3). From fig. 14 it can be seen that these requirements have been met [13]). The requirement that there should be a capacitor in parallel with the stray capacitances of each coil to earth (case 2), is met for coils which have one terminal connected to earth, if condition (1) is fulfilled. For those coils which have neither terminal earthed, the method normally used in high frequency measuring technique is employed. The coil is enclosed in a "live" i.e. not earthed screen connected to one of the coil terminals, *fig. 19*. The stray capacitance C_1 between the coil and the live screen can be regarded as part of the self-capacitance of the coil, and treated as for case (1). The capacitance between the live screen and earth can be regarded as being concentrated at the terminal to which the screen is connected and can therefore be treated as for case (3).

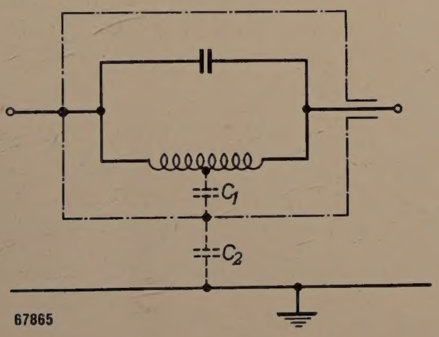

Fig. 19. When a coil is not earthed, the stray capacitances to earth are rendered innocuous by mounting the coil in a live screen connected to one of the coil terminals and dealing with the capacitances C_1 and C_2 separately as described in the text.

The filter panel [14])

Fig. 20 shows, in schematic form, the way in which the three filter networks are combined, together with amplifying stages and pads, to form the complete through supergroup filter.

The amplifying stages are identical, and each comprises a pair of EF 91 valves in parallel, in a circuit with negative feedback. Consequently, if one valve fails, the gain is reduced by only 0.5 dB; in that case the change of gain is uniform throughout the entire frequency band and the attenuation/frequency characteristic shown in fig. 15 is quite unaffected except for the change in datum value.

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[13]) The coil in fig. 14*b*, which appears to be an exception to this statement, can be regarded as an approximation to a negative capacitance, which was found to be theoretically necessary at this point.

[14]) The construction of the prototype filter, electrically designed in the Physical Research Laboratory at Eindhoven was undertaken by Dr. Latimer and Messrs. Jackson and Snelling of the Mullard Electronic Products Ltd. London.

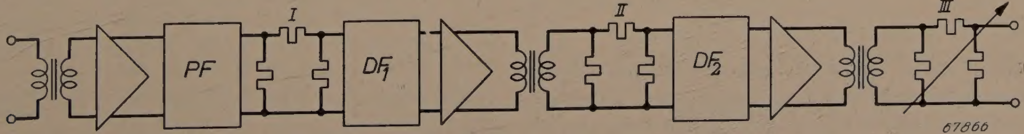

Fig. 20. Block diagram of the complete through supergroup filter. In addition to the two half filters DF_1 and DF_2, and the pilot stop filter *PF*, there are three amplifiers and three pads (*I,II,III*). The available gain of the amplifiers is such that the attenuation over the whole of the pass band can, if desired, be made negative. This satisfies a requirement of the specification, not previously mentioned, that it should be possible to adjust the attenuation in the pass band in steps of 1 dB from 7 ± 0.5 dB and from −2 ± 0.5 dB.

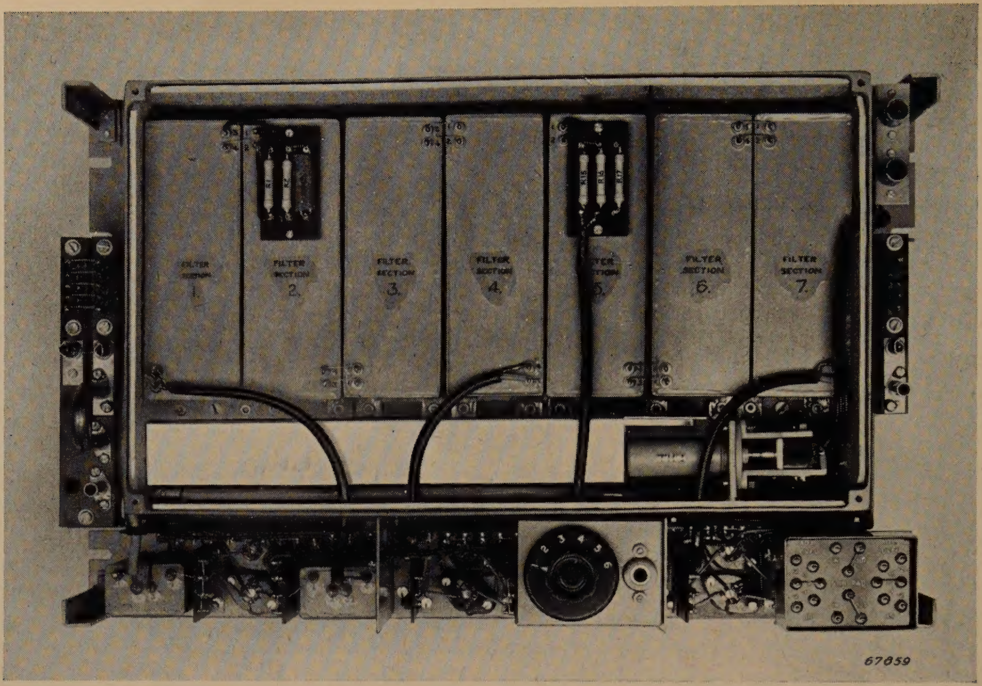

Fig. 21. The through supergroup filter, with dust cover and oven lid removed. The filter networks have been split up into seven convenient units, the mechanical construction and mounting of which is identical to that used for the filters of the 48-channel system described previously (see note [2]). Underneath these seven units can be seen the heater and thermostat assembly. Immediately below the oven are the amplifiers, transformers and pads. The variable pad (*III* in fig. 20) is at the extreme bottom right hand corner. The attenuation of this pad can be varied by altering the soldered connections.

The pad *I* has been introduced because both the pilot stop and band-pass filter sections have been calculated on the basis of termination with a purely ohmic resistance and should be terminated in practice with a fairly good approximation thereof, i.e. with a high return loss. Pad *II* has been introduced for the same reason, because the impedance looking back into the output terminals of the transformer of the preceding amplifier is not a sufficient approximation of an ohmic resistance. The total gain can be regulated over a range of 5 dB by means of the variable pad *III*.

Temperature variations cause variations in the attenuation/frequency characteristic. A variation of temperature from 10 °C to 40 °C causes a variation of attenuation of about 0.5 dB in the pass range of the filter, and unless the temperature were controlled, the limit of ± 0.125 dB allowed within a group would be exceeded. The filter networks are, therefore, enclosed in a simple thermostatically controlled oven, in which the temperature is kept at 45° ± 2° C.

In conclusion, *figs. 21* and *22* show the complete filter, with and without dust cover. Some features, visible on these figures, are mentioned in the subscripts.

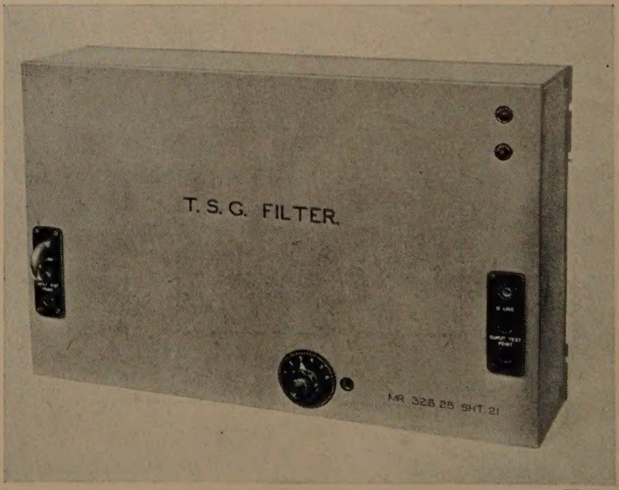

Fig. 22. Front view of the through supergroup filter, with dust cover in position. At the top right-hand corner, the two thermostat indicating lamps can be seen. One of these lamps lights up when the element is on and the other when it is off. The screen-grid current of any of the amplifying valves can be measured by selecting the particular valve required by means of the rotary switch at the bottom of the panel, and plugging a meter into the socket alongside the switch.

Summary. Coaxial carrier telephone systems are capable of transmitting 600 or 960 channels over two coaxial tubes (one "go" and one "return"). The carrier circuits are arranged in supergroups of 5 groups, each group comprising 12 channels.

In order to meet the needs of operational flexibility, facilities must be provided for the transit connection of one or more groups or supergroups at intermediate stations on a coaxial trunk network. In some cases it may even be necessary to connect all the channels through, except for one supergroup. One solution to this problem is to demodulate all the supergroups to the basic supergroup position of 312-552 kc/s. A through supergroup filter at this point permits transit working without further demodulation, and thus avoids the use of considerable additional modulation equipment. The specification which such a through supergroup filter has to meet is given, and is followed by a description of a prototype filter constructed using coils with Ferroxcube cores and mica capacitors. As a result of the very small losses of these components, and the application of the frequency parameter methods in the design of the filter, all the requirements of the specification have been met by a filter occupying a comparatively small space. (The use of quartz crystals for such a filter is not advisable on account of the risk of undesirable crystal resonances in the pass range.) The prototype through supergroup filter was built up from two identical band-pass filters and a supplementary pilot stop filter, separated by amplifiers and pads. The filter sections are housed in a temperature-controlled oven, to reduce the variation of attenuation with temperature.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS OF THE N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these papers not marked with an asterisk can be obtained free of charge upon application to the address printed on the back cover.

1957: C. J. Bouwkamp: On Sommerfeld's surface wave (Phys. Rev. **80**, 294, 1950, No. 2).

A critical discussion of some recent papers by Kahan and Eckart on Sommerfeld's surface wave in electromagnetic wave propagation over a plane earth.

1958: J. M. Stevels: The relation between the dielectric losses and the composition of glass (J. Soc. Glass Technology **34**, 80-100, 1950, No. 158).

Discussion of the theory of dielectric losses in glass. Losses are expected due to (1) relaxation phenomena caused by the network modifiers, (2) relaxation phenomena caased by the network itself and (3) vibration of the constituents of the network and of the network modifiers. The dependence of each kind of losses on the frequency and the temperature is discussed. The theoretical considerations are compared with the experimental results and the dependence of the various kinds of losses on the composition is discussed.

1959: F. A. Kröger and J. van den Boomgaard: The fluorescence of magnesium germanate activated by manganese (J. Electrochem. Soc. **97**, 377-382, 1950, No. 11)

Magnesium germanate activated by manganese may contain the activator in different valencies. When this valency is higher than two, fluorescence excited by ultraviolet occurs in seven narrow bands decaying with the same time constant. The variation of this decay constant and the intensity of fluorescence with temperature is explained by the assumption that the relative contribution of radiative and non-radiative transitions is determined by a Boltzmann equilibrium regulating the population of various excited states.

1960: J. A. Haringx: The rigidity of corrugated diaphragms (Appl. sci. Res. A **2**, 299-325, 1950, No. 4).

When the corrugated diaphragm is replaced by a fictitious flat plate of similar properties it is possible to derive a linear differential equation for the deflection. The coefficients of this equation, however, vary in a complicated way and its solution for the pressure-loaded diaphragm is given only for thick and for thin sheets separately. For thick sheets the profile of the corrugation appears to be inessential, whereas for thin sheets it is necessary to distinguish between trapezoidal, triangular and arc-shaped corrugations. By an obvious device the results for thick and for thin sheets are fitted together, so that the deflection can be determined also for the intermediate range of medium sheet thickness. The final results of the present calculation are compared with measurements carried out by others and are found to be in satisfactory agreement with the experiments.

It is to be remarked that, compared on the basis of small deflections, the introduction of corrugations into the sheet leads to a considerable increase of rigidity of the diaphragm. The prevailing assertion that the flat plate is more rigid than the corrugated diaphragm holds only for large deflections, because of the non-linearity between the load and the deflection of the flat plate.

1961: J. L. Snoek and E. J. Haes: Laboratory apparatus for the production of a steady flow of very pure hydrogen (Appl. sci. Res. A 2, 326-328, 1950, No. 4).

An apparatus is described, containing as a diffusor element a system of twelve thin-walled, nickel tubes in parallel, which, upon being heated to 950 °C and exposed to an excess pressure of about one atmosphere, yields a steady flow of about one quarter of a litre per minute of the purest hydrogen at atmospheric pressure. The apparatus is rugged and simple and has a lifetime of several hundreds of hours.

R 148: C. J. Bouwkamp: On Bethe's theory of diffraction by small holes (Philips Res. Rep. 5, 321-332, 1950, No. 5).

In this paper on the diffraction of electromagnetic waves by a small circular hole in an infinite plane conducting screen, Bethe derives the diffracted field from fictitious magnetic charges and currents in the hole. A careful examination of Bethe's theory reveals that the electric field should be discontinuous in the hole, which is inconsistent. The reason is that Bethe's expression for the magnetic current density is erroneous, as is shown in the present paper for the simplest possible case, viz. the diffraction of a plane-polarized electromagnetic wave impinging on the hole in the normal direction. Though the error in the current density does not invalidate Bethe's formulae for the distant field, the correct electric field in and near the hole differs appreciably from that predicted by Bethe.

R 149: J. L. Meijering: Segregation in regular ternary solutions, I (Philips Res. Rep. 5, 333-356, 1950, No. 5).

The segregation in ternary phases is examined from a thermodynamic point of view, the comments being confined to regular solutions. Pertinent information is already gained by analyzing the spinodal curves (boundaries between metastable and unstable concentration regions). The parameters a, b and c, defining the mutual energy interactions of the three component pairs, are varied independently of each other. Their different combinations are divided into eight characteristic categories. Taking the sequence $a \geqslant b \geqslant c$, closed miscibility gaps appear above $T = a/2R$ if $b - c > a > 0$, and above $T = 0$ if $\sqrt{-c} > \sqrt{-b} + \sqrt{-a}$. An expression is given for the ternary critical point in which such a gap vanishes on heating. Complicated spinodal curves are found when $b + c > a > 0$.

R 150: H. C. Hamaker and Th. Hehenkamp: Minimum-cost transformers and chokes, I (Philips Res. Rep. 5, 357-394, 1950, No. 5).

Three equations play a part in the theoretical design of a transformer: (1) the "price equation", specifying the price as a function of the geometrical dimensions, (2) the "power equation", expressing the apparent power in terms of the dimensions and of the magnetic-flux density and electric-current density, and (3) the "losse equation", giving the losses in terms of the same set of variables. Assuming the power and the losses prescribed, the problem is to find a minimum in the price as given by the first of these calculations, the other two serving as auxiliary conditions. This problem is solved and the solution is casted into a form ready for immediate application. The resulting tables with full instructions for use have been collected in the appendix. It turns out that the minimum-cost design depends on the ratio of power to losses, so that an increase of both in the same ratio does not alter the dimensions of the transformer, a conclusion unacceptable in practice. Frequently a design is not restricted by the losses, but rather by a permissible maximum of the temperature, or of the peak magnetic-flux density. It is discussed how such other characteristics can theoretically be brought into play. Readjustments lead to make-shift solutions which are adopted only by lack of a rigorous theory. Some light may be thrown on their practical value by a comparison with alternative designs in which transformer characteristics other than the losses have been prescribed.

R 151: P. Cornelius: Proposals and recommendations concerning definitions and units of electromagnetic quantities (modifications) (Philips Res. Rep. 5, 395-396, 1950, No. 5).

Next to the symbol $j = JV$ for the magnetic dipole moment (in Wb·m) the symbol $m = MV$ for the magnetic area moment (in A·m^2) is introduced and in analogy therewith, next to the symbol $p = PV$ (in C·m) for the electric dipole moment, a symbol $f = FV$ (in V·m^2) for the "electric area moment" (V volume in m^3, J magnetic polarisation in Wb/m^2, M magnetisation in A/m, P electric polarisation in C/m^2, F "electrisation" in V/m).